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TITLE: LECTURE 2: DC POWER SUPPLIES AND HARD-TUBE POWER
CONDITIONING SYSTEMS

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"POWER CONDITIONING TECHNOLOGY"

(Course Coordinator: W. Sarjeant)

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HIGH VOLTAGE/PULSE POWER TECHNOLOGY

GRADUATE COURSE EECS596

University of New Mexico

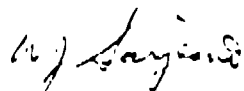
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PREFACE

With the recent increase in technological needs and the interest in the power conditioning arena, one of the problems facing workers in the field is the lack of texts or notes describing recent progress, particularly in the area of repetitive power conditioning. For this reason and because of expanding internal requirements, the University of New Mexico (UNM) and the Los Alamos Scientific Laboratory (LASL) have created a set of lecture notes based upon the graduate course taught recently at UNM. The objective of these notes is to create a record of many of the advances in the field since the last text in the field was published just after World War II. In this context, the lectures presented are oriented toward an introduction of the reader to each of the areas described and present sufficient background information to explain many of these advances. They are not intended to serve as design engineering notes, and thus the reader is referred to the references at the end of each lecture for detailed technical information in specific areas.

The preparation of these writings is a result of a considerable teamwork effort on the part of LASL and Sandia staff. In particular, Cathy Correll, in conjunction with Jo Ann Barnes and the rest of her efficient word processing staff, carried the major responsibility for preparation of the lectures while the lecturers did the proofreading and revisions. As course coordinator, it is a pleasure to acknowledge the strong support of Ray Gore, our E-Division Leader, and Shyam Gurbaxani who is the UNM Graduate Center Director, Los Alamos Campus.



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October 3, 1980

LECTURE 2

DC POWER SUPPLIES AND HARD-TUBE POWER CONDITIONING SYSTEMS

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DC POWER SUPPLIES AND HARD-TUBE POWER CONDITIONING SYSTEMS

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LECTURE 2:
DC POWER SUPPLIES AND HARD-TUBE POWER CONDITIONING SYSTEMS
by
W. J. Sarjeant

I. INTRODUCTION

In this lecture, we shall discuss some of the major characteristics of DC power supplies and hard-tube pulsers, along with many of their PCS limiting properties and several of the more powerful vacuum tubes that are currently available for larger hard-tube pulsers systems. The initial topic will be elementary DC power supplies (Fig. 1), operating with standard 60 Hz AC input power. The full-wave bridge design will be discussed in detail, illustrating the majority of the significant circuit parameters.

II. DC POWER SUPPLIES

If a simple DC power supply is snap started (suddenly energized to full voltage), and then the diodes, transformers and sometimes the filter choke are damaged, what has caused this damage is often a transient overvoltage effect rather than a short circuit on the output. These types of relatively low-frequency transients, which can cause damage and component breakdown in 60-400 Hz DC supplies, will be the subject of considerable discussion in this section. The standard 60 cycle, or 60 Hz, full wave bridge geometry (FWB) will serve as the design center, and the worst-case relations for all the design parameters applying to the diode, transformers, and filter design and selection, that apply all the way from full-wave doublers to 6-phase and 12-phase Y-delta DC power supply systems will be compared and related to these FWB parameters. For all of these classes of supplies, the same type of transient analysis will apply to

DC POWER SUPPLIES

THIS SECTION CONSIDERS ELEMENTARY DC POWER SUPPLIES OPERATING FROM 60 Hz AC POWER. THE FULL-WAVE BRIDGE DESIGN WILL SERVE TO ILLUSTRATE IMPORTANT CIRCUIT PARAMETERS. THROUGHOUT THE DISCUSSION, RELATIONS APPLICABLE TO OTHER GROUPS OF DC SUPPLIES WILL BE CORRELATED WITH THE FULL-WAVE BRIDGE (FWB) PARAMETERS.

CLASSES OF DC SUPPLIES ARE ILLUSTRATED IN THE "POWER SUPPLY" SECTION OF Terman (A HANDOUT).

ELEMENTARY TYPES AS SHOWN IN THE NEXT FIGURE

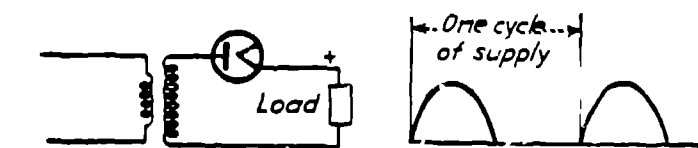
- A) HALF WAVE
- B) FULL WAVE, CENTER-TAP
- C) FULL-WAVE BRIDGE
- D) FULL-WAVE VOLTAGE DOUBLER
- E) VOLTAGE MULTIPLIER

FIG. 1

modeling the potential damage to components. A first-cut design of the power supply and the prerequisite damping networks can be obtained from this data and then, returning to the reference data that will be discussed, cost-benefit ratios for the supply design can be effected. In other words, what may come from the worst-case analysis here may be that the diode chosen has a higher current rating than that found necessary when a detailed circuit analysis has been executed. In this simplified, worst-case analysis, the design parameters determined will always err on the conservative side. One exception is the case in which the power line impedance is very low, potentially giving rise to enormous surge currents measuring hundreds of amps. Here for large, high-regulation, solid-state DC supplies of this type, it is necessary to obtain more accurate design data requiring a return to the fundamental relations in the references. The best source for most of the diode and filter design data is the RCA thyristor manual. In the back is an application note that is a summary of the work by Shade in the 1960s and it has almost all the design information needed to select a rectifier diode for the design. The other area discussed in the thyristor manual is how to execute relatively accurate ripple calculations when dealing with fairly precision supply requirements.

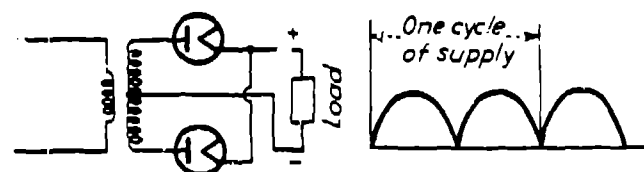
The five basic classes of elementary DC supplies are the half wave, the full-wave center tap, the full-wave bridge, the voltage doubler, and the voltage multiplier. These types of supplies are described in detail in the handout from Terman. Although it deals with tubes, this is probably one of the better analyses of the general characteristics of these types of supplies for ripple and the ratings the diodes must have. A summary of these different types of elementary supplies, A through E, is shown in Fig. 2.

- A. HALF WAVE: The load is pure resistive. There is no filter capacitor included, giving bursts of voltage, one for each positive going supply cycle .



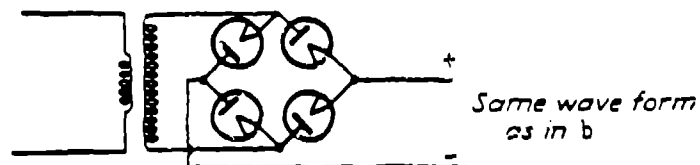
Half-wave (single-phase) output

(a) Half-wave rectifier circuit



Full-wave (two-phase) output

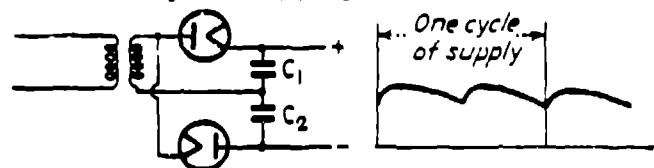
(b) Full-wave circuit with center-tapped transformer



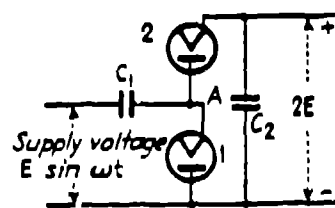
Same wave form as in b

(c) Full-wave bridge circuit

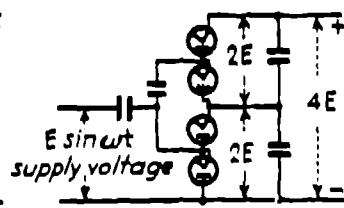
Voltage multiplying circuits



(d) Conventional voltage doubling circuit



(e) Cascade doubler



(f) Cascade quadrupler

- Terman

FIG. 2: ELEMENTARY DC POWER SUPPLIES

- B. FULL-WAVE WITH CENTER-TAP: There is a diode added to the bottom half of the transformer that gives a ripple frequency of 120 Hz, or twice the input frequency.
- C. FULL-WAVE BRIDGE: The main advantage of this supply is the high overall efficiency through full utilization of all the available transformer KVA or power rating.
- D. VOLTAGE MULTIPLYING CIRCUITS: These are basically charge stacking systems that can be used to provide very low currents at relatively high voltage (We will not discuss them in detail). At 60 Hz, they tend to be very difficult to design for currents above 10-15 milliamps, so they're basically very low average current supplies. For low current requirements, on the order of milliamperes, they're an inexpensive voltage multiplying technique. The slight disadvantage of using them is the relatively high RMS rating that the diodes and capacitors must have.
- E. COCKROFT-WALTON: This is the same type of voltage multiplier with a very high-frequency input. Then the reactances of the filter capacitors becomes very low, so that the amount of capacity needed to achieve very good regulation is small. Very large voltage multiplication factors with low ripples are obtainable.

FULL-WAVE BRIDGE DC SUPPLY - DETAILED ANALYSIS WITH CHOKE INPUT FILTER

Fig. 3 shows a sketch of the full-wave bridge DC supply. On the left side is the main power transformer, fed power through a set of line filters to keep transients from passing through in either direction, and a circuit breaker that is rated to trip on severe overloads. If the power supply is reasonably large, the contactor shown is employed to turn on the supply and it is generally energized by a latching circuit. The transformer feeds the diode full-wave bridge and the output voltage ripple

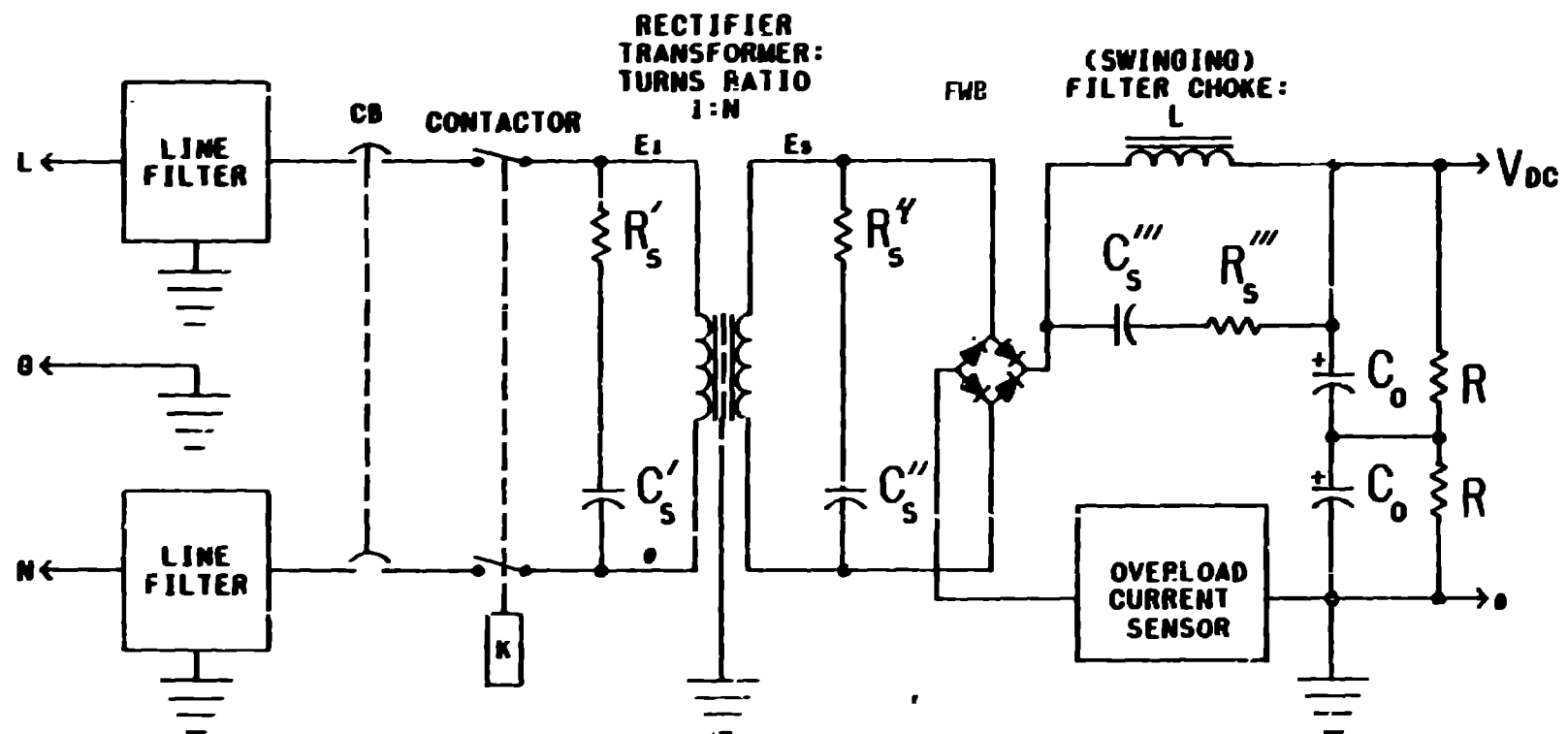


FIG. 3: FULL-WAVE BRIDGE DC POWER SUPPLY

is reduced by the LC filter. On a positive line cycle, current flows as shown by the arrows into the filter capacitors and back to the transformer. The transformer here is shown as having electrostatic shields, which are normally used in all sensitive circuitry or circuitry where there are relatively high voltages and transients, in order to shield the primary electrostatically from the secondary by a grounded metal structure. For single phase designs the shield cost is at most a few percent of the transformer cost, so there is no excuse not to ask the manufacturer to include them. Generally, for only a single shield, it is referred to the primary because of cost, since if a relatively high voltage is on the secondary, the added insulation for an electrostatic shield on the secondary is often expensive. The shield is generally a single layer of copper or aluminum foil over the primary, slit so the foil does not act as a shorted turn on the transformer, with a wire connecting the shield to the transformer case. That effectively puts an equipotential plane between the primary and the secondary as far as capacitive coupling is concerned, as no matter how the transformer is wound, there is some capacity from the primary to the secondary, and what is desired is that any current that the secondary tries to induce from capacitive coupling to the primary is carried off to the ground via the shield.

On the primary side of the power transformer in the DC supply, when the contactor is energized, the equivalent circuit here has the minimum inductance in the transformer, giving rise to large in-rush currents that must be taken into account in selecting circuit breaker and contactor surge ratings. In addition, a relatively large voltage transient can arise when the contactor is opened, caused by the collapse of the magnetic energy stored in the leakage and magnetizing inductances. The critical damping network, $R'_S - C'_S$, is employed here to dampen that transient when the contactor opens, particularly if one is unlucky enough to have it open at maximum current into the transformer. The damping network, $R''_S - C''_S$, on the secondary of the transformer is used for protection from both rapid turn-on (if turn-on is at the peak of the primary voltage this

network dampens the oscillations in the secondary loop), and for turn-off of the system when the diode FWB is then decoupled from the load loop as the transformer secondary voltage drops to zero. The diodes then no longer conduct current and are no longer pushing charge into the filter capacitor, so the secondary is just an RLC oscillator of R'_S , C'_S and the transformer leakage and magnetizing inductances (referred to the secondary). Selection of R'_S and C'_S values for near critical damping then controls the oscillation amplitude and duration. The same argument applies to damping transients across the filter choke. All these situations use an analogous damping technique.

The output of the supply in this case is shown as two filter capacitors in series, which indicates that, when the voltage is sufficiently high that a single capacitor is unavailable, it is necessary to stack a number of them in series and to use the voltage-grading resistors shown. A grading resistor is generally incorporated to pass on the order of 8 to 10 times the maximum leakage current through the capacitor at the maximum operating temperature of the supply. This ensures no more than a 15-20% imbalance in the voltage division across the 2 capacitors shown.

The overload current sensor in Fig. 2 serves to protect the diode bridge should the supply output be shorted to ground. It's timed to react within about 50-200 ms (typically this opening time can be achieved with fast electromechanical relays or vacuum interrupters). So if there is a fault on the output, the pulse current will energize the sensor, which then trips open the primary contactor (in this case, the important cost parameter is the number of cycles the diodes can carry the fault current before the contactor opens).

Next, let us discuss the characteristics of each of the individual sub-components in the DC supply. There is an efficiency advantage of a full-wave bridge over the center-tapped geometry because the full-wave bridge constantly loads the secondary of the transformer, whereas the center-tapped geometry unloads the transformer on alternate half cycles

when that particular portion of the winding is not conducting current. Also, during the recovery time of the diodes, it is possible for shock-excited oscillations to arise, damaging the transformer insulation or the reverse-biased diodes in the center-tapped geometry.

Fig. 4 shows a summary from Mlynar of various design terminologies for diodes and transformers that is presented for general information. The key parameters of concern in DC supply design are (* in Fig. 4) the repetitive peak reverse voltage applied to the diode, the worst-case asymmetrical fault current that the diode string can experience under start-up conditions or short-circuit faults in the load, and the KVA or power rating of the single-phase and three-phase power transformers. In general, most of the needed design information can be obtained from the table from Mlynar that is shown in Fig. 5. However, the question of transients and some of the simplifications possible in designing for low ripple aren't in there. These will be discussed presently.

In multiphase power supplies, each phase has a unique role to play. For very high powers (megawatts at hundreds of kV) the cheapest alternative is the six-phase series bridge (ref. 12-1-1-B in Fig. 5) because the voltage stresses on the transformer can be kept quite low (essentially half) keeping the corona level low. Interphase transformer power supplies in three-phase configurations can suffer from transient saturation effects when a pulse load is used (like a resonant-charging circuit or discharge networks).

Six-phase parallel operation, without interphase transformers is acceptable, except that fairly fast recovery diodes are needed to avoid one diode starting to conduct while the other one is trying to turn off, producing a high reactive kilovar loading of the secondary of the transformer. The reactive power that flows through the winding and the ohmage loss in the winding create significant transformer heating. The three-phase full-wave bridge is the best general-purpose three-phase supply, with either a Y or a Delta secondary configuration.

FIG. 4: TERMINOLOGY - DIODES AND TRANSFORMER

VOLTAGE

1. Design PRV - the arithmetic sum of the individual peak reverse voltages of all seriesed rectifiers; the absolute maximum reverse voltage that the assembly can withstand without catastrophic failure.
2. Rated PRV - the maximum reverse voltage normally expected based on the actual circuit operating conditions; the design peak minus any safety factor.
3. Transient PRV - the maximum allowable transient reverse voltage that might be expected under normal operating conditions.
- 4.* Blocking PRV - the maximum value of repetitive peak reverse voltage rating permitted by the manufacturer under stated conditions.
5. Output voltage - normally applies to the DC voltage output of a complete rectifier assembly.
6. AC voltage - the RMS input value, either line to line or line to neutral, of voltage applied to the rectifier assembly or bridge.

CURRENT

1. Rated current - the average value of DC current, generally the output current of a rectifier assembly; however, can be the per-leg average output current.
2. Short circuit current - the value of current that the output load could experience under continued short circuit conditions, usually limited by circuit impedance.
- 3.* Asymmetrical fault current - the peak value of current delivered to the output load caused by overshoot of the short circuit current; typically 1.5 to 1.8 times the short circuit current allowed by circuit conditions.

KVA

- 1.* Three-phase KVA using three single-phase transformers - system KVA is equal to three times the individual KVA of the single-phase transformers; however the system impedance is 3 times the individual single-phase impedance value.
2. Short circuit KVA - maximum KVA available under short circuit conditions.

FIG. 5: DC POWER SUPPLY DESIGN DATA FOR RESISTIVE LOADS

CONFIGURATION	SINGLE PHASE HALF-WAVE	SINGLE PHASE FULL-WAVE CENTER TAP	SINGLE PHASE FULL-WAVE BRIDGE	THREE PHASE HALF-WAVE WYE	THREE PHASE FULL-WAVE BRIDGE
SCHEMATIC					
SYMBOLIC NOTATION	1-1-1-H	2-1-1-C	4-1-1-B	3-1-1-Y	6-1-1-B
OUTPUT WAVEFORM					
PEAK REVERSE VOLTS PER RECTIFIER LEG (UNDER NO LOAD)	1414 V _{AC} <i>X 2.22</i>	2000 V _{AC}	1414 V _{AC}	245 V _{AC}	1414 V _{AC}
PEAK REVERSE VOLTS PER RECTIFIER LEG UNDER PURELY RESISTIVE OR INDUCTIVE LOAD ONLY	311 V _{DC}	310 V _{DC}	157 V _{DC}	200 V _{DC}	100 V _{DC}
RMS VOLTS OUTPUT RESISTIVE OR INDUCTIVE LOAD ONLY	187 V _{DC}	111 V _{DC}	111 V _{DC}	108 V _{DC}	100 V _{DC}
PEAK VOLTS OUTPUT RESISTIVE OR INDUCTIVE LOAD ONLY	314 V _{DC}	141 V _{DC}	157 V _{DC}	121 V _{DC}	105 V _{DC}
NO LOAD RMS AC VOLTAGE V _{AC} (ML)	222 V _{AC} (ML)	222 V _{AC} (ML)	111 V _{AC} (ML)	140 V _{AC} (ML)	74 V _{AC} (ML)
AVERAGE D.C. OUTPUT CURRENT PER RECTIFIER LEG	100 I _{DC}	5 I _{DC}	5 I _{DC}	333 I _{DC}	333 I _{DC}
RMS CURRENT PER RECTIFIER LEG	157 I _{DC}	707 I _{DC}	707 I _{DC}	577 I _{DC}	577 I _{DC}
I _{AC} RMS RMS CURRENT ON THE LINE BETWEEN THE TRANSFORMER AND THE RECTIFIER	157 I _{DC}	707 I _{DC}	100 I _{DC}	577 I _{DC}	541 I _{DC}
RATED POWER OF THE TRANSFORMER PRIMARY	247 I _{DC} V _{AC} (ML)	111 I _{DC} V _{AC} (ML)	111 I _{DC} V _{AC} (ML)	121 I _{DC} V _{AC} (ML)	105 I _{DC} V _{AC} (ML)
RATED POWER OF THE TRANSFORMER SECONDARY	39 I _{DC} V _{AC} (ML)	157 I _{DC} V _{AC} (ML)	157 I _{DC} V _{AC} (ML)	100 I _{DC} V _{AC} (ML)	100 I _{DC} V _{AC} (ML)
INTERPHASE TRANSFORMER RMS VOLTAGE (LINE TO NEUTRAL)	—	—	—	—	—
EQUIVALENT POWER RATING OF INTERPHASE TRANSFORMER	—	—	—	—	—

V_{AC} - RMS AC SECONDARY VOLTAGE
V_{DC} - AVERAGE D.C. OUTPUT VOLTAGE OF RECTIFIER $V_{DC} = \sqrt{2} V_{AC} (ML)$
I_{DC} - AVERAGE D.C. OUTPUT CURRENT OF RECTIFIER I_{DC} V_{AC} (ML) - IDEAL OUTPUT POWER OF THE RECTIFIER

THREE PHASE DOUBLE WYE	FOUR PHASE FULL-WAVE CENTER TAP	FOUR PHASE FULL-WAVE PARALLEL BRIDGE	FOUR PHASE FULL-WAVE SERIES BRIDGE	SIX PHASE STAR (THREE PHASE DIAMETRIC)	SIX PHASE PARALLEL BRIDGE WITHOUT INTERPHASE TRANSFORMER	SIX PHASE PARALLEL BRIDGE WITH INTERPHASE TRANSFORMER	SIX PHASE BRIDGE BRIDGE
6-1-1-Y	4-1-1-C	8-1-1-B	8-1-1-B	6-1-1-B	12-1-1-B	12-1-1-B	12-1-1-B
245 V _{DC}	2000 V _{DC}	1414 V _{DC}	1414 V _{DC}	2000 V _{DC}	130 V _{DC}	1414 V _{DC}	1414 V _{DC}
200 V _{DC}	222 V _{DC}	111 V _{DC}	707 V _{DC}	200 V _{DC}	101 V _{DC}	100 V _{DC}	541 V _{DC}
100 V _{DC}	100 V _{DC}	100 V _{DC}	100 V _{DC}	100 V _{DC}	100 V _{DC}	100 V _{DC}	100 V _{DC}
105 V _{DC}	111 V _{DC}	111 V _{DC}	111 V _{DC}	105 V _{DC}	105 V _{DC}	105 V _{DC}	105 V _{DC}
171 V _{DC} (ML)	157 V _{DC} (ML)	707 V _{DC} (ML)	500 V _{DC} (ML)	140 V _{DC} (ML)	74 V _{DC} (ML)	74 V _{DC} (ML)	37 V _{DC} (ML)
167 I _{DC}	25 I _{DC}	25 I _{DC}	5 I _{DC}	167 I _{DC}	167 I _{DC}	167 I _{DC}	333 I _{DC}
200 I _{DC}	50 I _{DC}	50 I _{DC}	707 I _{DC}	400 I _{DC}	400 I _{DC}	200 I _{DC}	577 I _{DC}
200 I _{DC}	50 I _{DC}	707 I _{DC}	100 I _{DC}	400 I _{DC}	577 I _{DC}	400 I _{DC}	541 I _{DC}
105 I _{DC} V _{AC} (ML)	111 I _{DC} V _{AC} (ML)	111 I _{DC} V _{AC} (ML)	111 I _{DC} V _{AC} (ML)	120 I _{DC} V _{AC} (ML)	101 I _{DC} V _{AC} (ML)	101 I _{DC} V _{AC} (ML)	101 I _{DC} V _{AC} (ML)
140 I _{DC} V _{AC} (ML)	157 I _{DC} V _{AC} (ML)	157 I _{DC} V _{AC} (ML)	100 I _{DC} V _{AC} (ML)	101 I _{DC} V _{AC} (ML)	143 I _{DC} V _{AC} (ML)	105 I _{DC} V _{AC} (ML)	100 I _{DC} V _{AC} (ML)
252 V _{DC} (ML)	—	—	—	—	—	405 V _{DC} (ML)	—
405 I _{DC} V _{AC} (ML)	—	—	—	—	—	—	—

Fig. 6 presents a tabulation from Terman for the choke-input DC power supply parameters, and Fig. 7 illustrates these rectifier circuits. In this case, there is a large inductor in series with the rectifier assembly, and one chooses this inductance to be sufficiently large so that at all times there is a current flow through that inductor. We'll see later that there is a critical value of the inductance, depending upon the minimum load current. If the current in the load becomes sufficiently small, a point is reached at which the inductor no longer conducts current continuously into the capacitance; there is thus no longer a unidirectional current flow through the circuit filter network so that the stored energy in the inductor can start impressing current into the filter capacitor if the net current flow is small enough, and the output voltage then rises to the peak voltage of the transformer. This is the turnover point in the regulation performance of the system, going from the choke input to capacitor input performance. This means that when a choke input supply is designed, there is a minimum load current that must flow through the output loop at all times. This is normally not a particularly difficult problem if one selects a type of choke whose inductance is a function of the current passing through it (swinging choke). The range in inductance from the minimum current to full-load current can normally exceed five to one, falling at the higher current as the choke core approaches saturation.

This summary, then, has indicated some of the salient aspects of single and multiphase DC power supply design. The tables allow selection of the RMS value of the transformer secondary voltage and current ratings. The maximum diode inverse voltage, for example, for the double three-phase series connection (Fig. 6) was 1.05 and for the single phase full-wave center-tapped connection it was 3.14. Analysis by Greenwood discloses that the peak overvoltage in any of these systems under stop-and-start conditions is about two. On rare occasions, peak overvoltages of three have been noticed. The usual engineering practice is to multiply diode inverse voltage requirements by two. This factor of two (a safety factor) almost guarantees that the device will not break

Rectifier Circuits

CHOKE INPUT DC POWER SUPPLY	Single- Phase Full- Wave Center- Tapped	Single Phase Full- Wave Bridge	Three- Phase Star	Three- Phase Broken Star	Six Phase Single Y	Double- three- Phase With Balance Coil	Double- three- Phase Series Conne- ction	Six Phase Star
	1b	1c	2a	2b	2c	2d	2e	2f
Voltage Regulations (d-c component of output voltage taken as 1.0):								
a. Rms value of transformer secondary voltage (per leg).....	1.11*	1.11	0.855	0.985	0.428	0.855	0.428	0.340
b. Maximum inverse voltage.....	1.14	1.57	2.09	2.09	1.05	2.42	1.05	2.39
c. Lowest frequency in rectifier output (f_r = frequency of power supply).....	2F	2F	3F	3F	6F	6F	6F	6F
d. Peak value of first three a-c components of rectifier output:								
Ripple frequency.....	0.667	0.667	0.250	0.250	0.057	0.057	0.057	0.057
Second harmonic of ripple frequency.....	0.133	0.133	0.057	0.057	0.014	0.014	0.014	0.014
Third harmonic of ripple frequency.....	0.057	0.057	0.025	0.025	0.005	0.006	0.006	0.006
e. Ripple peaks with reference to d-c axis:								
Positive peak.....	0.363	0.363	0.209	0.209	0.0472	0.0472	0.0472	0.0472
Negative peak.....	0.637	0.637	0.395	0.395	0.0528	0.0930	0.0930	0.0930
Current relations:								
f. $\frac{\text{Average current per anode}}{\text{Peak anode current}}$	0.500	0.500	0.333	0.333	0.333	0.333	0.333	0.167
g. $\frac{\text{Average current per anode}}{\text{Direct-current in load}}$	0.500	0.500	0.333	0.333	0.333	0.167	0.333	0.167
h. $\frac{\text{Peak current per anode}}{\text{Direct-current in load}}$	1.000	1.000	1.000	1.000	1.000	0.500	1.000	1.000
Transformer requirements (d-c output power = 1.0):								
i. Primary kva.....	1.11	1.11	1.21	1.05	1.05	1.05	1.05	1.28
j. Secondary kva.....	1.57	1.11	1.48	1.71	1.05	1.48	1.48	1.81
k. Average of primary and secondary kva....	1.34	1.11	1.35	1.38	1.05	1.26	1.26	1.55

NOTE: This table assumes that the input inductance is sufficiently large to maintain the output current of the rectifier substantially constant, and neglects the effects of voltage drop in the rectifier and the transformers.

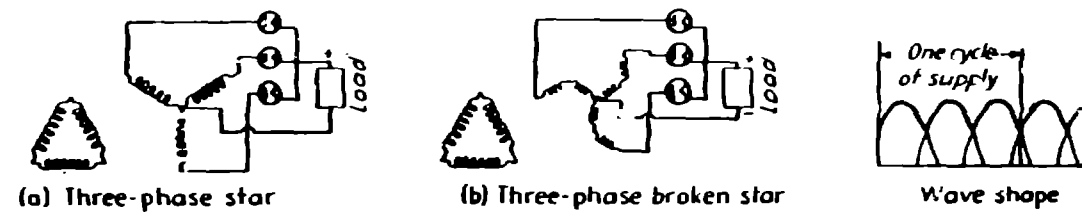
* Secondary voltage of one side of center tap.

The principal component of voltage across the balance coil has a frequency of 3F and a peak amplitude of 0.500. The peak balance coil voltage, including the smaller higher harmonics, is 0.605.

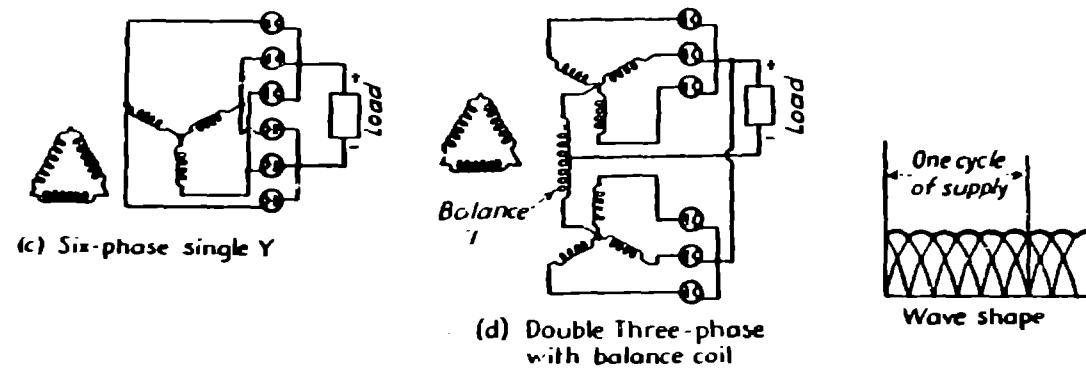
Refer to Fig. 2 for schematic diagrams of each rectifier circuit.

- Terman

FIG. 6: RECTIFIER CIRCUIT PARAMETERS FOR RESISTIVE LOADS OR VERY LARGE INPUT INDUCTANCE VALUES.



Circuits giving three-phase output wave



Circuits giving six-phase output wave

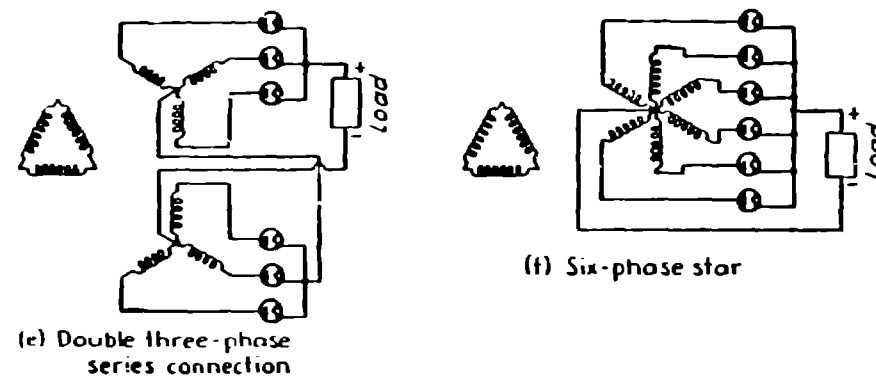


FIG. 7: RESISTIVE-LOAD OR CHOKE-INPUT RECTIFIER CIRCUITS

down. If, however, one deals with very high voltages, for example with 200-500 kV diodes, the ratio of cost to kV becomes larger and larger, making high kV ratings ever more expensive. In that particular case, the design is chosen to be as close to the safety margin as possible but, in general, not below 1.5 times the anticipated inverse voltage per leg is used. For example, for the above two cases, the inverse voltage ratings for the diode strings would be 1.6 and 4.7 times, respectively. In Fig. 6, Part d indicates that the ripple frequency amplitude components are expressed as a factor of the DC output voltage. For example, given a double three-phase series connection, the peak-to-peak ripple is .057 times the average output voltage. In Fig. 6, parts f to h, the peak diode current is compared to the direct-current in the load. The output voltage then rises to the peak of the AC input voltage. With the peak and average current through the diode known, one can go back to the RCA monograph and, after a brief analysis, determine the safety factors to apply and the diodes to choose. The last entries in Fig. 6, parts i-k, pertain to the transformer kVA relative to the DC output power as 1.0. Transformer kVAs up to a few hundred are not hard to buy, but diode costs rapidly escalate above 3A (average current).

Returning now to the full-wave bridge, its principal advantage is the superior transient suppression (Figs. 8 & 9) and more efficient use of transformer KVA. The center-tap design has a pair of solid-state diodes connected to a filter capacitor. During the transition when this diode becomes nonconducting, a small recombination charge current flows through it leading to a step in the recovery response of the diode. When that happens, this current shock excites this part of the transformer into oscillation. An R-C shunt network across the secondary will tend to dampen these oscillations. In the case of the full-wave bridge, the transformer is fully loaded for both cycles, minimizing the formation of such transients.

In the choke input case for the FWB (Ref. Fig. 6), if V_{DC} is the output DC voltage at I_{DC} , then if regulation is defined this way: the

FULL WAVE BRIDGE (FWB) POWER SUPPLY

THE MAIN ADVANTAGE OVER THE CENTER TAP DESIGN IS THE FAR SUPERIOR TRANSIENT SUPPRESSION OF FWB DESIGN DURING RECTIFIER COMMUTATION. ALSO, THE MAXIMUM INVERSE VOLTAGE IN THE FWB CASE IS ONE-HALF THE CENTER TAP DESIGN.

CHOKE INPUT

LET V_{DC} BE THE (FULL LOAD) OUTPUT DC VOLTAGE AT CURRENT I_{DC} . DEFINE % REGULATION $= \frac{V_{NL} - V_{DC}}{V_{NL}}$

FOR A CHOKE INPUT FILTER THE MINIMUM INDUCTANCE AT THE MINIMUM LOAD CURRENT IS: $\frac{\omega L}{R_L^{max}} \geq \frac{E_{ripple}}{V_{DC}}$ $\omega = 2\pi \times \text{ripple frequency}$
 $E_{ripple} = \text{peak ripple voltage}$

IF A LOWER CURRENT IS DRAWN, THEN THE CHOKE BECOMES INEFFECTIVE AND THE OUTPUT VOLTAGE RISES TO THE PEAK VOLTAGE OF THE TRANSFORMER SECONDARY.

FOR FWB DESIGN

$$\frac{\omega L}{R_L^{max}} > 0.7 \quad \therefore \quad \frac{L}{R_L^{max}} > 10^{-3} \quad \text{AND} \quad R_L = \frac{V_{DC}}{I_{DC}}$$

RIPPLE

$$\text{PEAK VALUE} \approx \frac{0.7 V_{DC}}{\omega^2 LC}$$

$$\text{CF FOR RC FILTER} \approx \frac{0.7 V_{DC}}{\omega RC}$$

$$\text{WHERE } \omega = 2\pi (2 \times 60)$$

FIG. 8

SINCE: $\frac{R_L^{\max}}{L} \leq 10^{-3}$

$$E_{\text{output}}^{\text{ripple}} \approx 10^{-6} \cdot \frac{V_{\text{DC}}}{(LC)} \quad \text{FOR EXAMPLE, } C = \frac{C_0}{2} \text{ IN FIG. 3}$$

COMPARE TO THREE-PHASE FWB WHERE:

$$\frac{R_L^{\max}}{L} < 40,000 \quad \text{AND} \quad E_{\text{ripple}}^{\text{output}} \approx 10^{-8} \cdot \frac{V_{\text{DC}}}{(LC)}$$

THUS THE DESIRED RIPPLE AND LOAD RANGE DEFINES THE VALUE OF L.
SWINGING CHOKES OFFER IMPROVED REGULATION FOR A LARGE RANGE IN

R_L VALUES: A $\frac{\Delta L}{L}$ OF FIVE TIMES IS OBTAINABLE.

THE VOLTAGE RATING OF CHOKE = $2 \times V_{\text{DC}}$.

TRANSFORMER PARAMETERS IN THE FWB SUPPLY

- | | | |
|--|---|---------------------------------------|
| 1. $V_{\text{RMS}} = 1.11 V_{\text{DC}}$ | } | NEGLECTING CORE AND
WINDING LOSSES |
| 2. $\text{KVA} = 1.11 \text{ DC OUTPUT POWER}$ | | |

DIODE PARAMETERS IN THE FWB SUPPLY

$$\text{PRV} \geq 2 V_{\text{DC}} \quad (\times 2 \text{ SAFETY FACTOR})$$

$$I_{\text{AV}} = \frac{1}{2} I_{\text{DC}} \quad \text{AND } I_{\text{PK}} = I_{\text{DC}}$$

R:
—

SELECT TO BE LESS THAN 10% OF THE LEAKAGE RESISTANCE OF CAPACITOR C_0 AT THE MAXIMUM TEMPERATURE. THIS ENSURES AN EQUAL DIVISION OF VOLTAGE TO WITHIN $\pm 20\%$.

FIG. 9

no-load voltage minus the DC voltage over the no-load voltage, then the choke input filter minimum inductance for a given R_L is:

$$\frac{\omega L}{R_L^{\max}} \geq \frac{E_{\text{ripple}}}{V_{\text{DC}}}$$

The angular frequency, ω , (2π times the ripple frequency) times L (at the current V_{DC}/R_L^{\max}) over the maximum value of the load resistor has then to be larger than the peak ripple voltage over the DC voltage. If a lower current is drawn, there is a point at which the choke is no longer effective (Fig. 10). A measure of the ripple on the output of the filter is obtained from Fig. 6. If instead of an LC filter an RC filter is used, the ripple at ω , the angular ripple frequency, would be considerably larger. From the calculation of the specified ripple required one can work back and calculate the value of the LC product, under the constraint of the relations in Fig. 9. That capacity, C , is the total capacity on the output of the LC filter, but in the case of the full-wave bridge we had two capacitors, each of value C_0 , in series so that:

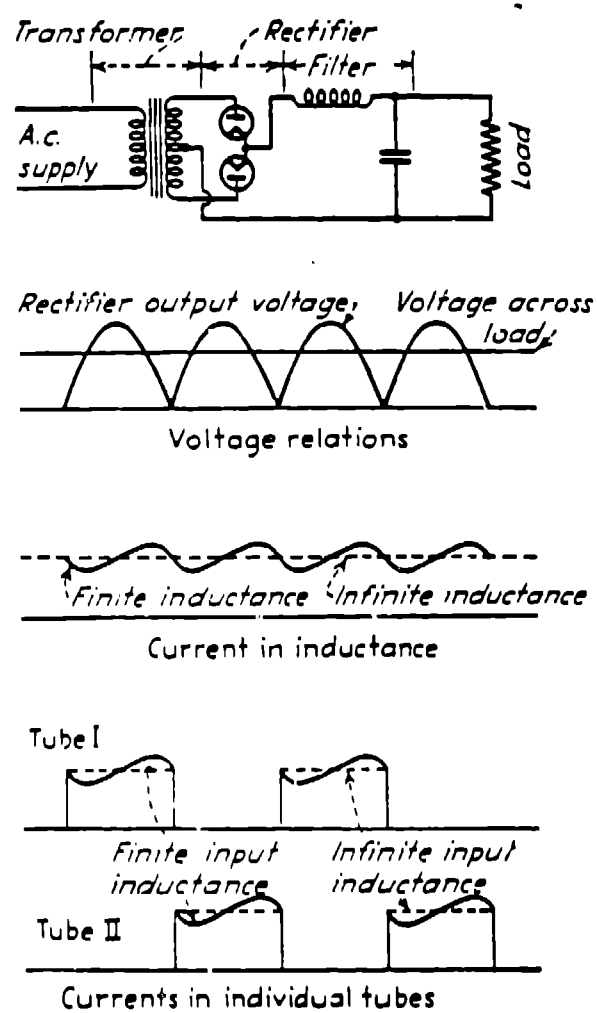
$$C = \frac{C_0}{2}$$

If a broad range in output current is required, one can derive L_{\max} from the amount of tolerable ripple, the minimum load current, and the swinging choke inductance ranges available. Given L and C , all the remaining design parameters in the system can be calculated.

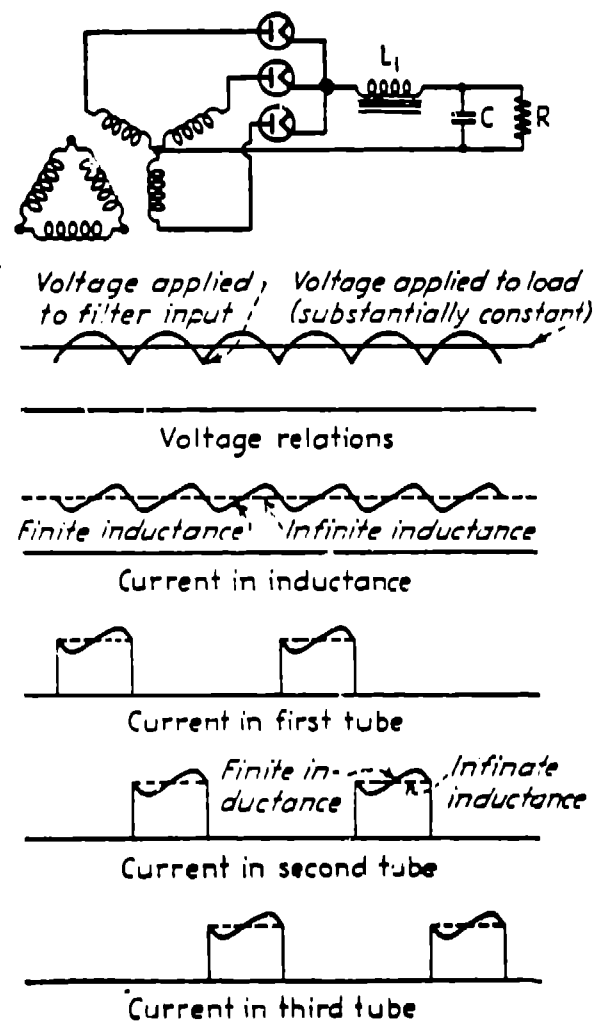
In the full-wave bridge, the peak output ripple is:

$$10^{-6} \cdot \frac{V_{\text{DC}}}{LC}$$

In this case, V_{DC} is the actual DC output voltage. This is to be compared to the three-phase power supply, where the increased ripple frequency results in a factor of 100 times reduction in the ripple. That's one of the real advantages of three-phase over single-phase operation, other than regulation. It's either easier to filter to a given



(a) Circuit of rectifier and filter - two-phase case



(b) Circuit of rectifier and filter - three-phase case

- Terman

FIG. 10: EFFECT OF A FINITE INPUT INDUCTANCE ON CURRENT WAVESHAPES

ripple level or, if a much smaller ripple is desired, the same size of inductors and capacitors can be used to gain several orders of magnitude reduction in the ripple. The major point about a swinging choke is that if there is a large range of load resistance to deal with, a choke whose inductance increases as the current decreases can keep the above relationships relatively constant. Now, a typical range in inductance is 5:1. Normally, in the design of filter choke one assumes that at turn-on there is zero voltage on the output, while peak volts are on the input. The insulation stress on the choke at 60 Hz is normally taken as twice that peak voltage -- the safety margin then being a factor of two. Recall that the safety margin in the diodes is also a factor of two. The rationale for that is in the economics of manufacture. It's relatively cost-effective to design, up to voltages of 100 kV, with such an insulation and diode safety factor. The damping network then controls the transient voltages to a factor of twice the peak voltage. This all fits in rather nicely with the design factor of two for the insulation level and defines the peak overvoltages to be experienced under worst-case conditions.

Where substantial high-voltage insulation is necessary in high-voltage chokes, the choke can then cost as much as or more than a modest-sized power transformer. A typical choke, for example, that might cost several hundred dollars to buy in an open mounting configuration could cost a factor of 2 to 3 more if hermetically sealed with a pair of high-voltage feed-through bushings. So the optimal solution is generally to put all power supply components in one container, fill it with transformer oil, and put a top on it.

A shunt resistance, R , is needed across each of the filter capacitors, C_o . Normally if R is chosen to be 10% of the leakage resistance at the maximum operating temperature, then the percent variation of the voltage across C_o under all conditions caused by temperature change is $\pm 20\%$. The capacitor heats up and cools down (from the environment and ripple current heating), and the leakage

resistance can vary over a wide range. If the resistors were not there, it's conceivable that one of the capacitors could have a leakage resistance 5-6 times smaller than the other, causing the voltage rating of the latter to be dangerously exceeded.

The third option in regulation control for the choke input DC supply when no-load to full-load operation is needed is to use a bleeder resistor. If the power supply is going to operate effectively from no load to some average current, this resistance can be chosen to draw sufficient current to ensure that the swinging choke is operating within the constraint of the relations in Fig. 9. The bleeder resistor can be combined with the shunt resistance, R , to simplify construction.

We now address the question of voltage division across the diode stacks as shown in Fig. 11. Consider a number of diodes in series (say N) and a very sharp transient applied across them (the typical capacitances of each diode being 100-300 pF). The transient wavefront appears across the input diodes in proportion to the diode to distributed (C_p) capacitance ratio. It's quite conceivable that a particular diode experiences a transient voltage well in excess of its reverse voltage holdoff capability. What normally happens in a high-voltage stack is that this diode burns out, and with the next pulse the next diode burns out, then the next, and so forth all the way down the stack. The voltage across each of those diodes can be kept close to the peak voltage divided by N , by using relatively large capacitors across the diode (capacitors, in fact, that are much larger than any of the distributed capacitances or diode capacitances in the circuit). Under these conditions, the voltage grading becomes quite uniform and the transient voltage jumps across each of the divides are quite small. Fig. 11 (2-3) is an illustration of what happens when no shunting components are included in the system. There are very sharp spikes at the beginning and the ending of the commutation intervals. At the end of the interval where the diode becomes non-conducting, the energy stored in the diode and distributed capacitances has to discharge. Where does it go? Since the peak voltage

FIG. 11: TRANSIENT VOLTAGE DIVISION ACROSS DIODE STACK ELEMENTS

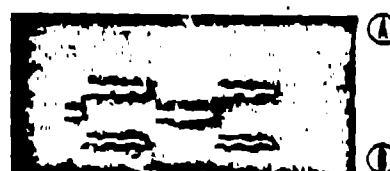
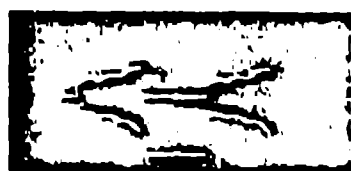
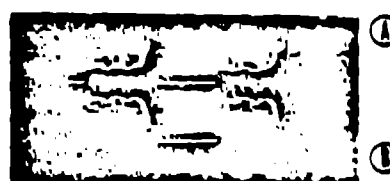
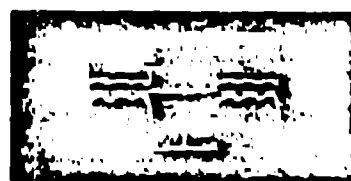
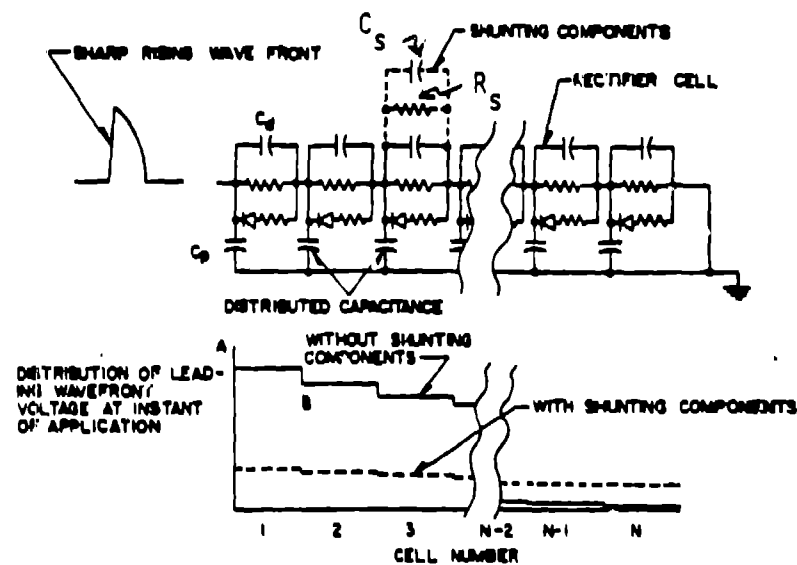


FIG. 2-5 Voltage Division with a Shunt Capacitor Equal to 100 Times the Junction Capacitance

FIG. 2-6 Voltage Division with a Shunt Resistor Equal to the Reverse Resistance

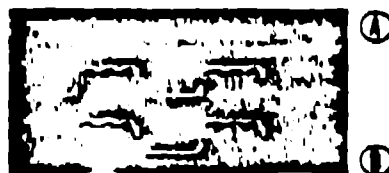


FIG. 2-7 Voltage Division with a Shunt Resistor Equal to the Reverse Resistance and a Shunt Capacitor Equal to 100 Times the Junction Capacitance

① Diode A

② Diode B

can then exceed the diode PRV, the diode is forced back into conduction in the reverse direction, avalanches, and the current pulse travels up through the stack. That little current bump can cause significant damage by destroying the diodes during the avalanche breakdown. If an appropriate shunt resistor-capacitor network is inserted across each diode, the input and output become virtually identical. The shunt capacitances are normally in the range of 1-3 nanofarads. Most of the modest-sized diode structures (1 to 5A) can be fabricated up to the several-hundred-kV level.

Experimental diode stacks can be built in the lab quite inexpensively. Avalanche-protected diodes are unnecessary. An avalanche-protected diode is one that behaves like a Zener diode if a voltage in excess of its avalanche-rated inverse voltage is applied. That's supposed to protect it. There are a number of manufacturers who do not make RC-compensated diode stacks. Virtually everything they make is avalanche protected. The diode leakage is matched to within a few percent and many of these diodes are stacked in series. A number of people who are using them end up incorporating RC compensation resistors and capacitors across each of those avalanche stacks. The point here is that the RC compensated stack guarantees equal voltage division even for very fast transients, particularly as experienced in laser systems.

Fig. 12 shows several high-voltage assemblies made by Westinghouse illustrating individually RC diode compensation and the heat sink is a large, thick aluminum plate between each of the diode assemblies. At the top is an illustration of one of the capacitor-resistor shunt assemblies for each of the 35A diode elements. Westinghouse does make one now that uses this geometry but has a fully screened heat-sink arrangement so that they can be used in circuits where there are very steep transients on the leading edge. The reason that's a problem is that if a very fast transient is applied that rises more quickly than the propagation time of the stack, large transient overvoltages can occur and LC resonances can be shock excited causing diode elements to avalanche and burn out. If the

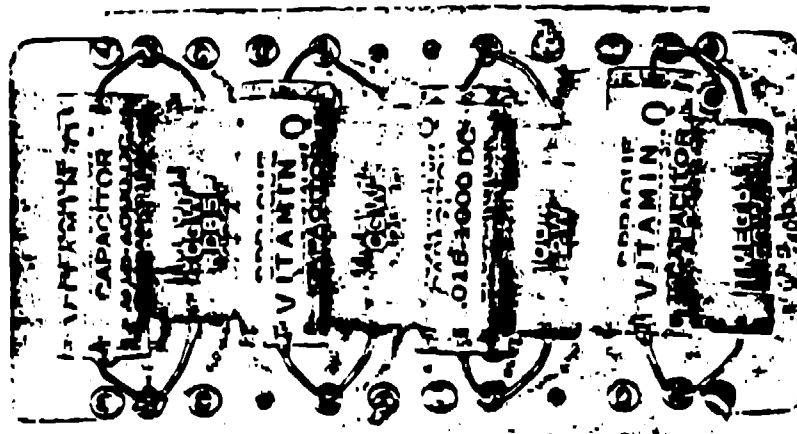


FIG. 1-19b Back side of a module showing churning components for four 25 ampere rectifiers

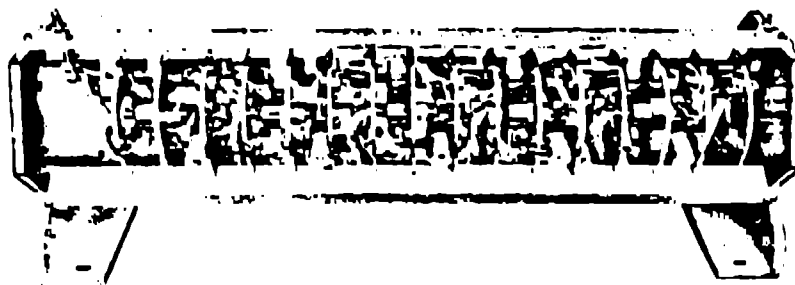


FIG. 1-20 70 ampere high voltage assembly utilizing circular aluminum cooling fins



FIG. 12: HIGH VOLTAGE, RC-COMPENSATED DIODE STACKS

diode assembly is encased in a metal box with only a little slot in it, one side being the input side and the other side being the output side of the diode, two equipotential planes are formed. What happens inside is screened from the outside world. These devices are made up to 150 A at the 500 kV level.

Most of the small stacks that can be bought (average currents up to 1A) generally use avalanche diodes. The solid state world being what it is, it's fairly straightforward at this current level; automated machines are available that select out and grade according to reverse current leakage factors and take the diodes in each group and stack them all together. To use one of these stacks in a fast circuit, the best thing to do is put a capacitive shunt across the total circuit, whose value can be estimated from the number of diode elements in the stack and the assumption that a shunt network of 3 nanofarads would be correct for each individual diode.

Individual RC-compensated diodes can be bought from IRC and EDI, among others. There's a curve in Fig. 13 giving an indication of diode reliability as a function of junction temperature, showing a very strong argument, then, in favor of buying as much average current capability as one can afford in a diode stack for maximum reliability. There's been a tendency in the high-voltage supply business over the years to use 1N4000 series diodes for all 1 A power supply stacks. Unfortunately the junction temperatures tend to run relatively high and the percent failure goes up. For a given known probability of failure of the device the weakest link in the system tends to be a solid-state diode running at a high junction temperature and the second weakest link tends to be transient suppression. If you compensate them properly and carefully, the devices still most likely to fail at high temperatures in the system, statistically, are the solid-state diodes. They're very good, but the hotter they run the more problems they're likely to have. The power dissipated in the diode in watts (Fig. 13) is given in the manufacturer's data as a function of average current. The case temperature also goes up

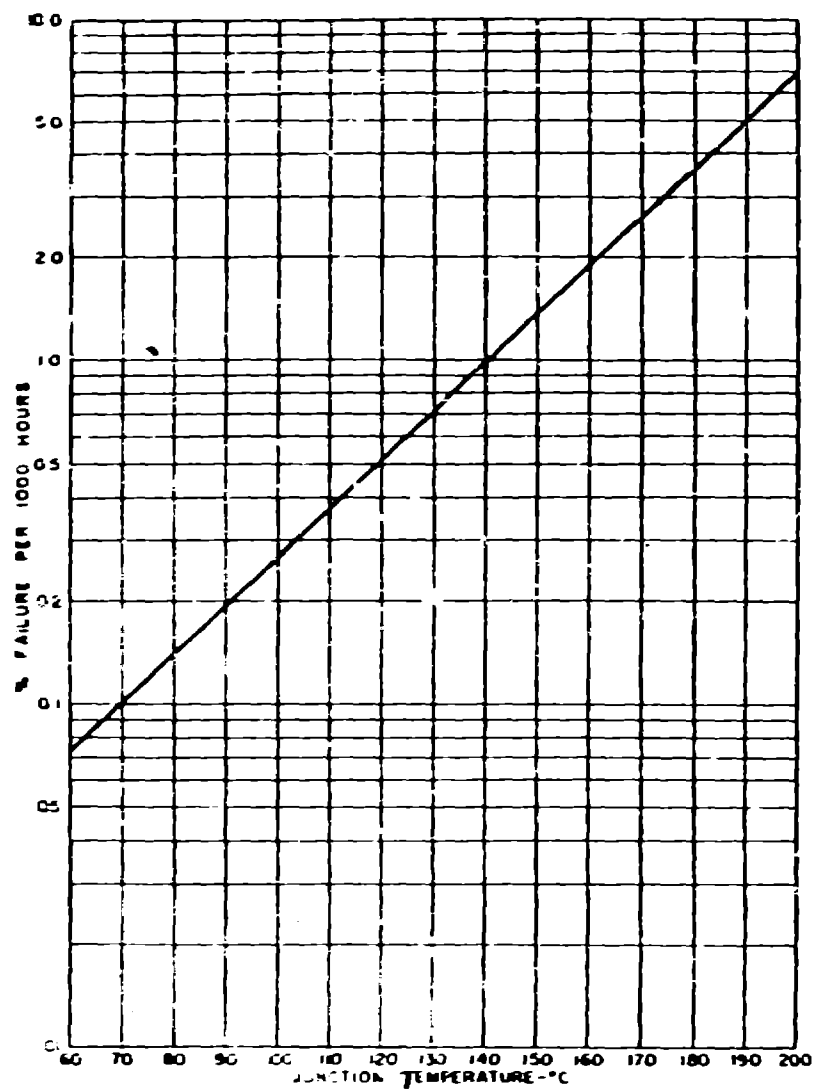
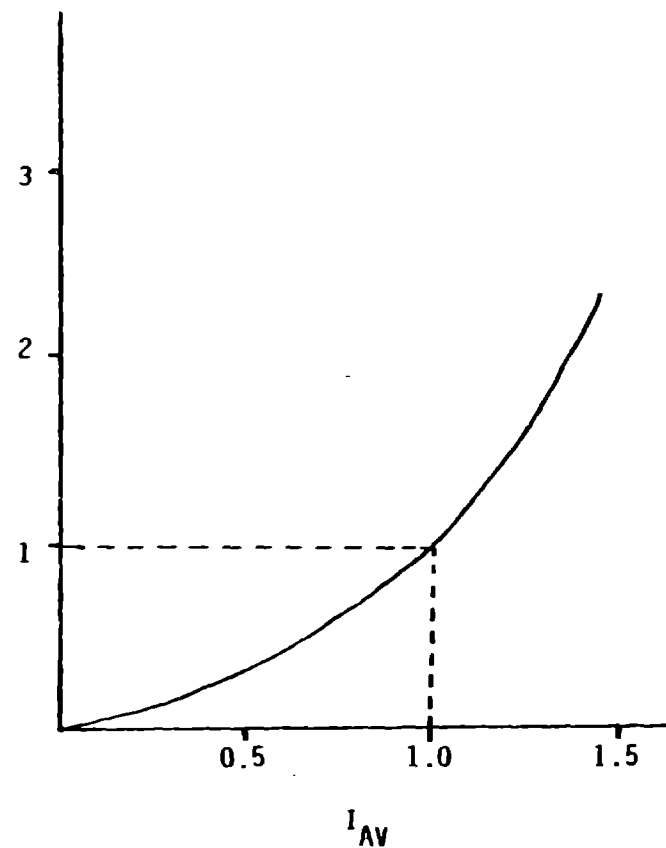


FIG. 13-1 Junction Temperature - true Percent Failure per 1000 Hours

FIG. 13: SOLID STATE RECTIFIER RELIABILITY



$$\theta_{JC} = 5^{\circ}\text{C/watt}$$

Given T_A , we can obtain T_j

$$T_j = (5 + T_A) / (1 \text{ watt})$$

for T_j at one amp I_{AV}

so many degrees for every watt of average power dissipated. That number is available from the manufacturer. Then, to specify a certain failure probability, Fig. 13 gives the junction temperature, and one can then calculate the corresponding case temperature, defining what the heat sink thermal resistance ($^{\circ}\text{C}/\text{watt}$) rise above ambient has to be for that given case and ambient temperature. RCA has a set of notes out in their audio amplifier applications brochure that does this calculation rather nicely. This is a representative diode curve in Fig. 13 for a 10 A silicon diode that Westinghouse made in the 1960s. It might be a lot better today. There is a reliability problem if the junction temperatures become too high, and if it doesn't cost a great deal, it's advantageous from a reliability point of view to keep the junction temperature down. When it begins to cost a lot, it is necessary to make the heat sink capacity larger, reducing the junction temperature.

The manufacturer will generally provide his lifetime expectancy at least at the 99.5% confidence level. (There are some other applicable military standards for diodes.) Operating at the rated junction temperature, and therefore at a specific average current for the given thermal resistance, for a failure rate of 0.5%, only five of a thousand of these devices will then fail in a thousand hours. That assumes that the distribution is normal. Now, there is another way to make considerable improvement to this reliability, and that's to fold into that the number of hours the device is actually on vs on-off cycling. That's another component of reliability previously unmentioned. There is a report put out by the Boeing Corp. that provides additional information in this area. Heretofore, it was assumed that the power supply will work for so many hours and the diode will be on and under constant load. Cycling it on and off can buy lifetime. If it's on for a short period compared to thermal equilibrium time (adiabatic operation) and off for a couple of days, this curve is weighted enabling the use of smaller devices. Experimentally, this appears to work, although the data base is sparse for high-power devices.

CAPACITOR INPUT FWB DC POWER SUPPLY

The other interesting power supply is the small FWB built in the lab. It serves as a full-wave bridge power supply, but uses a capacitor input instead of an inductor input filter at the output of the FWB. The capacitor charges to the peak secondary voltage. Fig. 14 illustrates the characteristics of the full-wave capacitor-input power supply. In the FWB case, the ripple frequency is doubled. There are some changes in the transformer requirements and the diode requirements. What will be discussed here are worst cases based upon very good transformer performance -- the lowest core losses and the lowest winding losses practical -- and then some relations can be derived that will give an upper bound to device requirements. This will allow determination, for a diode with specific characteristics and a transformer with a given set of characteristics, whether or not any of the components are going to burn out. In the case of a transformer, the kVA or power requirement is as reported for the choke input plus the three loss terms (which the manufacturers provide) shown in Fig. 15. When dealing with a relatively low-current supply (an amp or two at a few hundred volts), these losses matter little. These factors really become important when dealing with high currents or fairly high-voltage supplies where the transformers must be specified carefully for the best cost/performance tradeoff. During iteration with the manufacturer there will be a cost tradeoff. He can provide low copper losses that tend to improve regulation, but it will cost considerably more. As a rule of thumb, 2-10% is a reasonable regulation range to consider. Asking for 1% is very expensive for large supplies. These are manufacturing restrictions, not fundamental engineering or physics restrictions. Now the diode losses were estimated from the curve on the reliability figure. The diode conducts such an average current and dissipates so many watts, and those are lost watts. Those must be compensated for by increasing the power capability of the power transformer. From the diode curves in the RCA handbook the worst-case RMS current rating for the transformer is 5 times I_{DC} , and I_{PK} is 40 I_{DC} . This indicates that one should go back and have another close look, because RMS current capability in a transformer has a large copper cost for a given percentage loss and this power loss must be folded into the total kVA lost in the transformer.

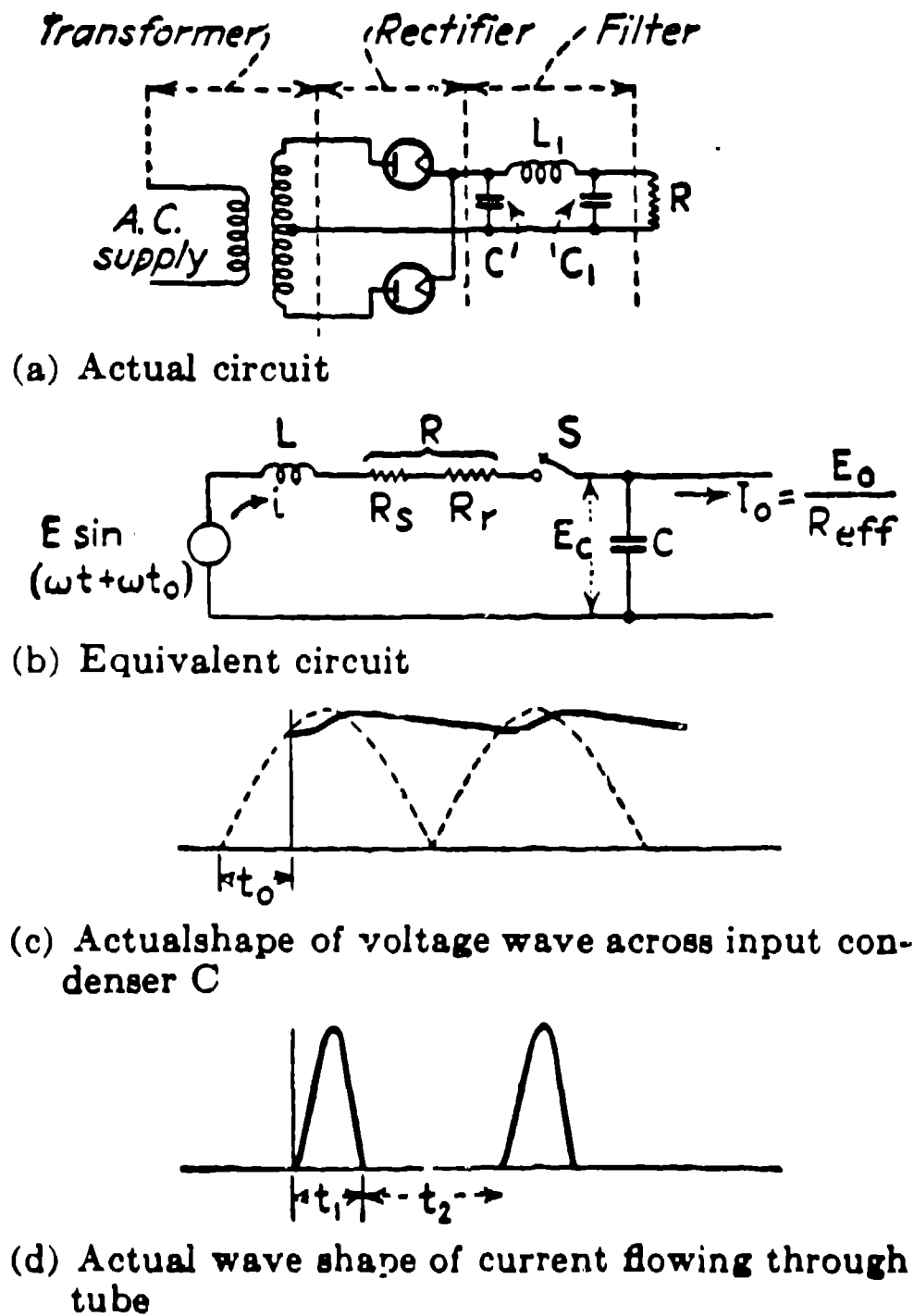


FIG. 14: ACTUAL AND EQUIVALENT CIRCUITS OF A FULL-WAVE CAPACITOR-INPUT POWER SUPPLY.

In designing supplies where the transformer is to be bought, it is suggested to spend a considerable amount of time interfacing with the transformer manufacturer. He can build into his transformer a certain amount of reactive component that doesn't represent a real power loss but one that does allow the limiting of the fault current in the secondary when it is shorted. Effectively this puts an inductor in series with the transformer primary as a current-limiting element, physically, in the way the transformer is built. In addition, transients can be significantly reduced by the different ways in which the transformer windings are configured. A great deal depends on whether one winds the primary directly on the secondary, beside it, or on another leg of the core. By interfacing with the manufacturer considerable money can often be saved for a given performance level requirement. A typical transformer cost saving on a large supply can be a factor of 2 or 3, and for a transformer costing in the \$50k region, this can be worthwhile.

The most expensive approach is to prepare a set of design specifications, send them out for quotes, and wait for everybody to write back with the costs. It costs time and money. The more effective route is to call the various people in the transformer-manufacturing field and talk with them. It might take a morning. When all that information is available, a more cost-effective design is achieved. In most cases the cost of the telephone calls is more than recuperated. This is particularly true in cases of having to build very high kVA supplies at modest voltages (say, 5-10 kV).

Diode costs can be staggering for the larger kVA high-current supplies. The parameter estimates given in Fig. 15 are a worst case and are inappropriate for very large (hundreds of kilowatts to megawatts) supplies. The analysis done in the RCA note advocates the selection of a PRV twice the peak voltage. This factor of 2 times is a reasonable approach to use for up to 100 kV supplies. Above 100 kV it's probably too expensive and 1.3 to 1.5 times can be used, provided very careful fault-mode analysis modeling is executed. Westinghouse assures you with

FIG. 15: FVB POWER SUPPLY

CAPACITOR INPUT

- SHORT OUT INPUT INDUCTOR L IN FIG. 3.
- $C = \frac{C_0}{2}$ NOW IS CHARGED TO THE PEAK SECONDARY VOLTAGE.

(1) TRANSFORMER PARAMETERS

- A. $V_{RMS} = 0.7 V_{DC}$
- B. $KVA = 1.11 \times P_{DC} + \text{WINDING LOSSES} + \text{CORE LOSSES}$
+ DIODE LOSSES.

CALCULATE: $I_{RMS} \approx 5 I_{DC}$ } FROM THE DIODE CURVES THE
 $I_{PK} \approx 40 I_{DC}$ } POWER LOSS IS OBTAINED.
 (THIS CAN BE ESTIMATED AS
 ≈ 1 WATT PER 2-A PEAK.)

(2) DIODE PARAMETERS

- A. $PRV = 2\sqrt{2} V_{RMS}$ WITH A SAFETY FACTOR OF 2
 - AT VERY HIGH VOLTAGES THIS IS USUALLY REDUCED TO:
 $PRV = 1.5\sqrt{2} V_{RMS}$.
- B. $I_{RMS} = 5 I_{DC}$ $I_{PK} = 40 I_{DC}$
- C. THE RIPPLE SPECIFIED DETERMINES THE FILTERING NEEDED
 WHICH THEN DEFINES C. REGULATION DEFINES THE ALLOWABLE
 SERIES CURRENT LIMITING RESISTANCE:
 FOR FVB: GOOD REGULATION AND LOW (≈20%) RIPPLE DEMANDS
 $v \omega C R_L = 50$ $v = 2$ $F = 60 \text{ Hz}$

NOTE: FOR HALF-WAVE $v = 1$, VOLTAGE DOUBLER $v = 1/2$;
 THE FOLLOWING APPLIES IN GENERAL:

THE RIPPLE RANGE IS 2% (FVB) TO 5% (FW DOUBLER).

FIG. 15A

31A

- CAPACITOR RMS RIPPLE CURRENT $\approx 5 I_{DC}$
- AT 120 HZ RIPPLE FREQUENCY FOR FWB
- AT 60 HZ RIPPLE FREQUENCY FOR FW DOUBLER.

90% confidence that 1.5 is sufficient PRV margin if a snubber is placed in the transformer primary and the secondary, as well as the filter choke. The worst-case RMS current rating of the diode should be 5 times the average current and the peak current is 40 times the average current (Fig. 16). This can then be compared to detailed model predictions (using, for example, NET-2) in order to reduce the current requirements of the components to a safe minimum value.

To obtain a degree of ripple reduction to 2%, find C from Fig. 15. Then the regulation-allowable voltage drop in the diode stack depends upon how much equivalent series resistance is present in each diode leg. Selecting, for 1A DC, a 40 A peak diode, the output is faulted and the peak current is obtained from the model (Fig. 16 and 17). For example, with a hundred ohm load and a 1 A supply, faulting the output for a perfect transformer, diode, and zero input line impedance, there would be infinite fault current. There is, however, some internal resistance in the supply. What this does is if the diode has a 40 A rating this says that for a 100 ohm load the internal supply resistance should be 100 over 40, which is 2.5 ohms. With a 2.5 ohm equivalent resistance inside the power supply, shorting the output of the power supply will keep the fault current to 40 A. Normally the transformer manufacturer can tell you the equivalent resistance of the transformer. This resistance is normally the DC copper resistance referred to the secondary of the transformer. There is also a small contribution from the diode stack. In this particular case for a 2.5 ohm internal resistance, faulting the output using a 40 A diode, the circuit breaker would open or the fuse would blow, and the diode would survive.

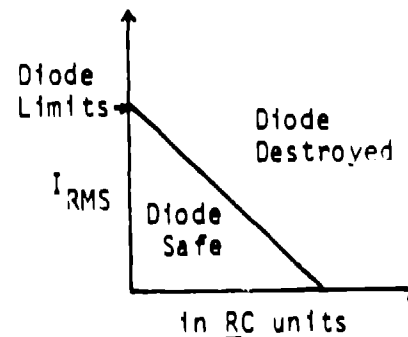
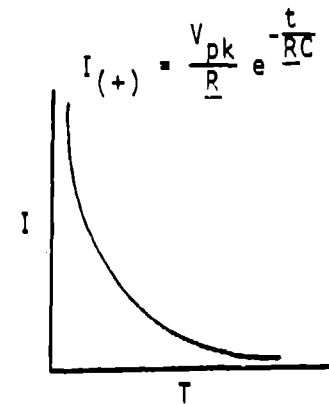
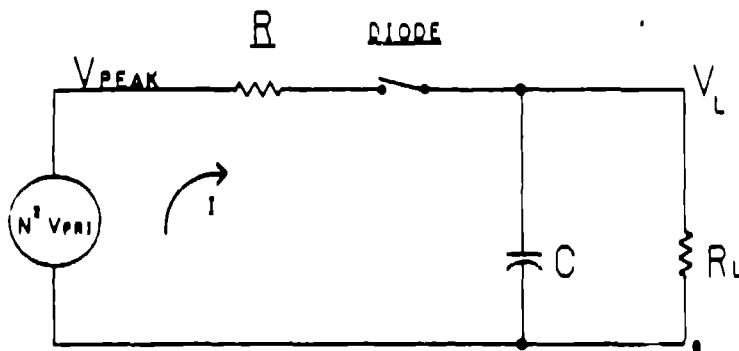
These calculations can become more sophisticated for three-phase or more multi-phase supplies or for complicated, very low ripple systems with a very high-degree of regulation. In the case of the full-wave bridge, if $\nu CR_L = 50$ is chosen, and $\nu=2$ in this case, for good regulation the ripple is also low ($\approx 2\%$). In the half-wave supply $\nu=1$, and for the case of a voltage doubler it is $1/2$. This relationship allows the calculation

FIG. 16

NOW THE DIODE SURGE RATINGS CAN BE ASSESSED:

LET \underline{R} BE THE TOTAL SERIES RESISTANCE =
WINDING + EQUIVALENT DIODE RESISTANCE + RESISTIVE
COMPONENT OF FILTER CAPACITOR REACTANCE.

$\underline{R} \ C = \text{SURGE TIME CONSTANT}$



IF \underline{R} IS NOT KNOWN THEN THE ITERATIVE TECHNIQUE MUST BE
USED: ASSUME $\underline{R} \approx 0.01 R_L$ (A WORST CASE). THEN, KNOWING FROM
RCA APPLICATION NOTES:

$$I_{RMS} \cdot (\underline{R} \cdot C) = 0.7 \cdot V_{PEAK} \cdot C$$

CALCULATE I_{RMS} AND CHECK ON DIODE GRAPH. IF ABOVE THE LIMIT
LINE, INCREASE \underline{R} . FOR A WORST-CASE ESTIMATE FOR $\omega C R_L = 50$,
 $\underline{R} = 0.01 R_L$ $I_{RMS} = 5 I_{AV}$ $I_{PK} = 40 I_{AV}$ (GENERALLY APPLICABLE)

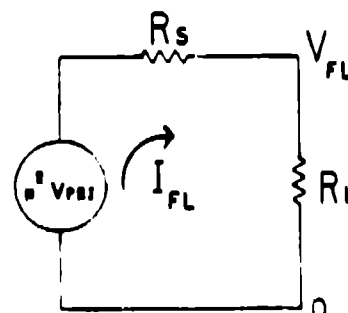
FIG. 17: FAULT CURRENT AT TURN-ON OR OUTPUT SHORT CIRCUIT

FOR A TRANSFORMER:

$$\% \text{ REGULATION} = \frac{V_{NL} - V_{FL}}{V_{FL}}$$

$$\text{SINCE } V_{NL} - V_{FL} = I_{FL} R_S = \frac{V_{FL}}{R_L} R_S$$

$$\text{THEN } R_S = (\% \text{ REG}) R_L$$



WHERE THE TRANSFORMER IS LOADED WITH A PURE RESISTANCE
SO THAT SECONDARY POWER IS EQUAL TO DC POWER DESIRED
PLUS LOSSES.

NOTE: GENERALLY $R_S \approx \underline{R}$ EXCEPT FOR LOW-POWER, HIGH-VOLTAGE
UNITS.

FAULT CURRENT:

FOR THE WHOLE SUPPLY, MLYNAR HAS SHOWN:

$$I_{\text{PEAK}} \approx \frac{I_{DC} \times 1.8}{(\% \text{ REG})} \quad \text{ASYMMETRIC FAULT CURRENT}$$

- AVAILABLE ON TURN-ON INTO A SHORT CIRCUIT
AT PEAK LINE VOLTAGE

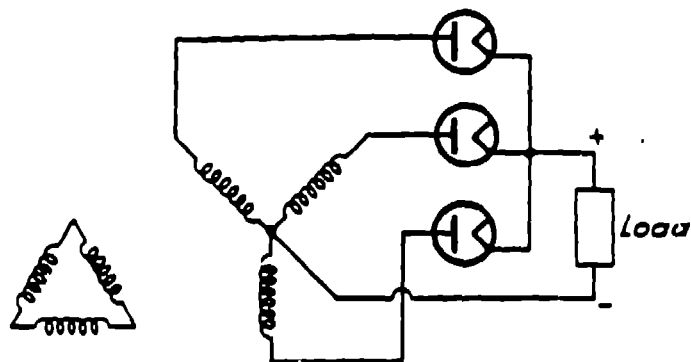
of C and determination of the ripple range. The ripple range is 2-5% (peak-to-peak), in going from the FWB to voltage doubler geometries. In attempts to reduce the amount of ripple further, the RCA curves reveal that peak current begins to rise enormously posing a clear trade-off situation. Should you put another LC filter in the supply? The answer is if one designs to remain near the fault current levels of Fig. 16 for these three classes of supplies, it's almost always more cost effective to put another LC filter in the circuit. In fact, on very high-voltage supplies, the additional filters can be inserted into the negative line. This is a very inexpensive way to make a low-ripple high-voltage supply with relatively inexpensive filter chokes. Most such chokes are rated for 1000 V RMS. The inexpensive way to protect them is to use a small spark gap (e. g. Centralab "Gap-Cap") that costs less than one dollar. It can be joined to a current-limiting resistor if desired. These gaps can handle about one joule of energy. For lab applications they're good for a couple of hundred faults. They are not quite as effective as a nonlinear varistor that has a softer damping curve (so that as the voltage across it is increased, it gradually increases the current) but they are inexpensive. To stack 4-5 chokes like this in series (which, although not recommended, will work) the choke cores are floated and the core is tied to the next lower choke output lead (if two or more of them are in series) and the Gap Cap placed across each set of choke leads.

During a fault in the output the chokes tend to develop the peak secondary voltage across them as they try to act as current regulators. The "Gap-Cap" shorts, essentially bypassing the fault current through the limiting internal resistance of the supply. There is another type of "Gap-Cap" that has a 0.01 microfarad capacitor inside it. These were extremely common in the days when radios had tubes where they would be used as protectors across the output transformer, particularly in radios that had remote extension speakers. Thyrites and metal oxide varistors are much better, but also far more expensive. For general laboratory applications, "Gap-Caps" (Centralab) are most effective.

. These relations in Fig. 15 allow one to select a value for C in the capacitor input case. Keeping $v_w C R_L = 50$, in this ripple range of 2-5%, will cause the RMS and peak currents to obey the relations in Fig. 16. That's true for all types of supplies except in multi-phase units. In general, in multi-phase supplies, these would be very conservative design ratings. Additional design information is available in Mlynar and Terman.

Let us next show the effects of leakage inductance with reference to Fig. 18. Leakage inductance just means that, between the primary and the secondary of the transformer, not all of the primary magnetic flux is coupled to the secondary. The loss arises primarily from geometrical factors. A small equivalent inductance is put in series with the AC driving voltage in the transformer equivalent circuit (Fig. 18). There is the resistance of the diode, R_D , and that of the winding R_S , of the transformer. Consider each as an ideal diode switch closing once every $1/180$ of a second. During each diode cycle, a large current pulse flows and puts a voltage on the load. Each diode is then sequentially reverse-biased and it's at these points during turn-on and turn-off that transient oscillations in the system can arise. The diode has to recover, it has stored charge in it, and the equivalent of an LRC generator in the supply circuit can then arise. The details of this are complicated and we will not address them. Practically, all that is required is to add damping networks to the primary and secondary of the transformer. If the circuit does oscillate, nowhere in this voltage ringing will the voltage exceed twice the AC peak value from the transformer.

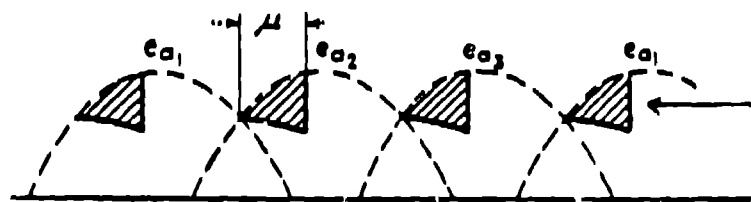
As capacitors become large, they can have a significant RMS ripple currents passing through them. In the case of the full-wave bridge the RMS current is about $5 \cdot I_{DC}$ at 120 Hz. It's just a point to be checked in the capacitor specifications to ensure that the capacitor will tolerate



(a) Three-phase half-wave rectifier circuit



(b) Current of individual anodes



(c) Voltage waves

Spike shock
excites resonanc
in filter choke
and power trans-
former.

FIG. 18
EFFECTS OF LEAKAGE INDUCTANCE ON OUTPUT VOLTAGE

that amount of current heating. In the case of the surge ratings, this aspect is covered in more detail in the RCA notes. For a given filter capacitance C and total internal resistance R of the power supply, the peak voltage instantaneously applied to this series RC network, through the diodes is N times the primary voltage of the transformer. Assuming an ideal diode, at the peak voltage the current flow is through an RC type of network. R in this case is this R . If the R is known, the peak current can be calculated, and the proper diode selected. If the current is not known, one may estimate for practical transformers, assuming that R_{eq} is 1% or R_L (quite a good assumption) and, calculating from the application notes, this means the transformer regulation is 1%, which is a firm minimum number, (Fig. 16) so that:

$$I_{rms} \cdot R \cdot C = 0.7 \cdot V_{peak} \cdot C$$

I_{rms} is calculated and checked against the diode curves available from the manufacturer. (Time is shown in RC units in the insert in Fig. 16.) If the number on the left side is above the limit line, the diode will not survive. What this means basically is that if the calculated RMS current is above the limit line the diode will melt at turn-on. This is really an energy limit above which the chip melts.

Percent regulation has already been defined (Fig. 17). The voltage difference between the no-load and the full-load voltage is the full-load current times R_s , and since:

$$I_{FL} = \frac{V_{FL}}{R_L}$$

then the internal supply resistance, R_s is:

$$R_s = (\% \text{ Regulation}) \cdot R_L$$

For a given percentage regulation the equivalent resistance can thus be estimated.

In this case, generally the equivalent loss resistive R_s is the DC (copper) resistive loss in the transformer plus the core loss in the transformer. The exception to this consideration arises in very large polyphase transformers with many amps through the diodes where asymmetric effects in the multiphase transformer can predominate. In this case, the peak current is the DC current divided by the percent regulation. There's an asymmetric fault current factor ratio of 1.8 to be applied for the peak fault current as discussed in Mlynar. This is the peak current available when the whole supply is turned on to a short circuit at peak line voltage. In conjunction with the transformer regulation and the average DC current, that will then indicate the absolute peak current to be expected, so far as the transformer or diode are concerned.

The last part of this lecture will discuss damping networks, what they do, transformer-equivalent circuits, oil insulation, and an introduction to hard-tube pulsers.

QUESTION: Is there a general text concerning high-voltage insulation and measurement techniques?

"High Voltage Engineering," by E. Kuffel and M. Abdullah (Pergamon Press 1970) covers many interesting areas such as these. It's probably one of the best practical books published on spark breakdown, ionization, decay processes, electric breakdown of gases, and a most important area -- measurement of high voltages. It introduces the reader to the literature of applied measurement techniques and forms a very valuable resource book. It actually covers practical areas, such as switch triggering techniques, many types of high-voltage generators, and the measurement section is comprehensive, starting at electrostatic volt meters and finishing with compensated probes.

TRANSIENT DAMPING

Throughout all pulse-power systems there are potential transient problems. It's usually the transient phenomena that cause device

breakdowns. Rarely in a well-engineered system, where these transients are under control, will the components fail. Where systems usually fail is when somebody turns on the power supply without the output load, or with the output load shorted, a transient is generated, a large peak voltage is generated in some series-resonant circuit and a capacitor, choke, diode, or switch is lost. What will be discussed in this section is the problem of damping networks (Fig. 19) to control these transients. Essentially the problem is one of voltage damping in the series resonant circuit. When the power is initially applied, there is some inductance, L , there's a surge resistance, R , and the filter capacitance, C . If the power sine-wave is applied at $t = 0$, the voltages will build up in the circuit until the equilibrium value is reached, generally requiring 10 to 30 periods at 60 Hz (i.e., ≈ 170 to ≈ 500 ms).

For the purposes of our discussion, the fundamental concern is with the absolute maximum peak-to-peak secondary voltages across each high-voltage component, not with their growth or decay. Given that, the next question is how to ensure that this voltage can be kept within controlled limits so that the diodes, inductors, resistors, or capacitors don't break down.

For a full-wave bridge, when the power supply is turned on, the magnetic core of the transformer must be magnetized: the amount of current needed to do so is called the magnetizing current. The area inside the B-H curve is related to the core loss per cycle (it takes one full period to retrace the curve). The magnetizing current is essentially the amount of current applied that generates the magnetic field for zero secondary load power. In general terms, we can call it I_E . It is required to start the motion on this curve. The smaller this curve area, the smaller the core losses will be. In traversing the curve, say counter-clockwise, work is being done in moving the orientation of the ferromagnetic domains of the material as the coercive force, $H(I)$, varies. The thermal losses arising from this motion (core losses) change as the frequency (usually drastically increasing) increases, and with the core material.

DAMPING NETWORKS

FOR THE SINGLE-PHASE FWB DC SUPPLY:

WHEN THE SUPPLY IS TURNED OFF AT THE PEAK OF THE LINE VOLTAGE AND THE LOAD IS NOT CONNECTED, THE PRIMARY CURRENT THROUGH THE MAGNETIZING INDUCTANCE " L_E " RAPIDLY COLLAPSES, GENERATING A HIGH VOLTAGE:

$$V_{pri}^{peak} \approx \left(\frac{L_E}{C_{shunt}^p} \right)^{\frac{1}{2}} I_{magnetizing}^{peak} \quad \text{:FROM CONSERVATION OF ENERGY}$$

UNLESS C_{shunt}^p IS SUFFICIENTLY LARGE, V_{pri}^{peak} CAN RISE TO LARGE VALUES, CAUSING PRIMARY INSULATION BREAKDOWN. TO CONTROL THIS, ADDITIONAL SHUNT CAPACITANCE IS ADDED AND, IF NECESSARY, SOME RESISTANCE IS PLACED IN SERIES WITH THE ADDED CAPACITANCE TO PROVIDE RAPID DAMPING OF THE TRANSIENT.

FIGURE 19

When the power supply is turned on, current flows in the transformer primary even if the secondary is not connected. To measure I_E an ammeter is put in series with the primary of the power transformer, the meter is shunted, and with the secondary open, power is applied. After a few seconds the shunt is opened. The average value is measured with the milliammeter and that can be converted to the RMS equivalent. Given that, the energy stored in the equivalent magnetizing inductance L_E can be measured in the following way: for an ideal transformer with a magnetizing inductance L_E , then the shunt damping capacitance C_s can be determined from conservation of energy

$$\frac{1}{2} C_s V_{pk}^2 = \frac{1}{2} L_E I_{pk}^2$$

As before, V_{pk} is set equal to twice the peak line voltage and L_E obtained from $V_{pri} = \omega L_E I_{pri}$. Then, since this is an oscillatory RLC circuit, this energy $\frac{1}{2} L_E I_{pk}^2$ is dumped into the capacitor and oscillates repeatedly. This calculation can then give an estimate of the peak voltage that can be expected. Depending upon C_s and L_E , V_{pk} on the secondary can exceed the power supply peak voltage by a factor of 3-4. To dampen this voltage to twice the AC peak voltage applied, we will show later on that the:

$$Q = \frac{\omega L_E}{R_s} = 2$$

Now C_s and R_s in series with C_s are determined.

This magnetizing energy is not lost. It's in the iron, causing magnetic energy to be stored in the lumped equivalent inductance of the transformer primary. A current meter put into the circuit will measure a current flowing through the system with no load on the secondary. The transformer there is thus not ideal. If the current is measured, the value of L_E with no applied secondary load is obtained. It may not be a real inductor, but L_E is represented as a lumped equivalent inductance.

The problem, depending upon the transformer parameters, is that there can be significant transient voltages developed well above the line voltage. What one wants to do is to have a way of damping this large peak, irrespective of initial conditions. The same sort of argument applies to the damping of the transients across the choke in the filter (Fig. 20). If this is an ideal transformer, it has a turns ratio of one to N so that if you apply one volt AC to the input, N volts AC are at the output. This is what everybody would like to be the case, but alas the equivalent circuit is more complex and is sketched in Fig. 21. There's a magnetizing inductance, $L_{\text{magnetizing}}^P = L_E$, and a shunt resistance

$R_{\text{core loss}}^P$ representing the core loss (the work done in moving the ferro-magnetic domains around on the B-H curve). This series resistance, R_{winding}^P is equal essentially to the resistance of the copper primary wire. The leakage inductance, L_{leakage}^P , which is shown lumped in the primary, represents essentially the percentage of the magnetic flux generated in a primary that is not fully coupled to the secondary. A rough number for that leakage inductance is 1-5% of the primary inductance. The transformer manufacturer can provide a more accurate number for a given design.

The equivalent elements on the secondary side are complimentary. One small element that can become important is the primary to secondary capacity C_{ps} . It's the real physical capacity from primary to secondary that one would measure with a capacitance bridge. In most circuits it's a relatively small number. The only time it becomes sufficiently large to be a problem is in multilayer, interleaved (multifilar) transformer designs popular in wideband pulse and audio transformers. For conventional pulse transformers in the microsecond region, it's preferable not to wind them with these interleaved structures to minimize C_{ps} and reduce the coupling of transients through the transformer.

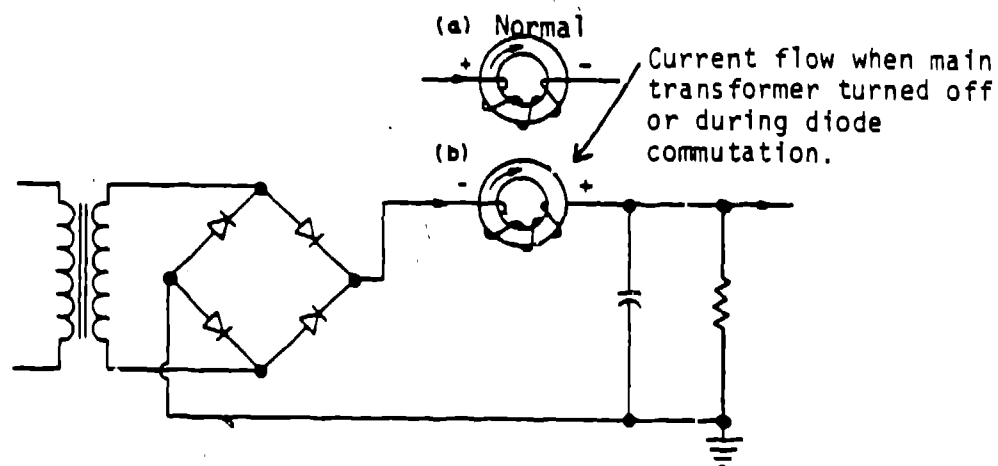
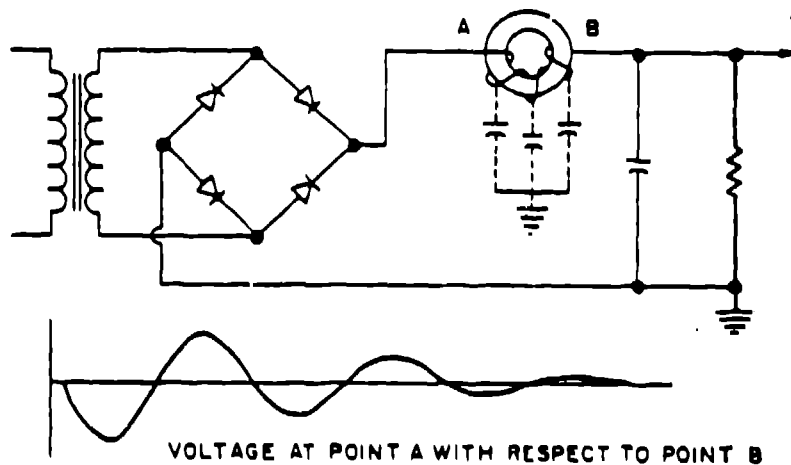


FIG. 7-50 Swinging Choke

The main factor in this arrangement is the distributed capacitance which is present from the coil to ground. This key parameter causes an oscillatory effect of this regenerative voltage and will produce a voltage of reverse polarity across the rectifier assemblies after the rectifiers have commutated the forward bias pulse. Refer to Figure 7-51.



- Mlynar

FIG. 20: TRANSIENT VOLTAGES GENERATED ACROSS THE INPUT INDUCTOR DURING COMMUTATION OF THE DIODES.

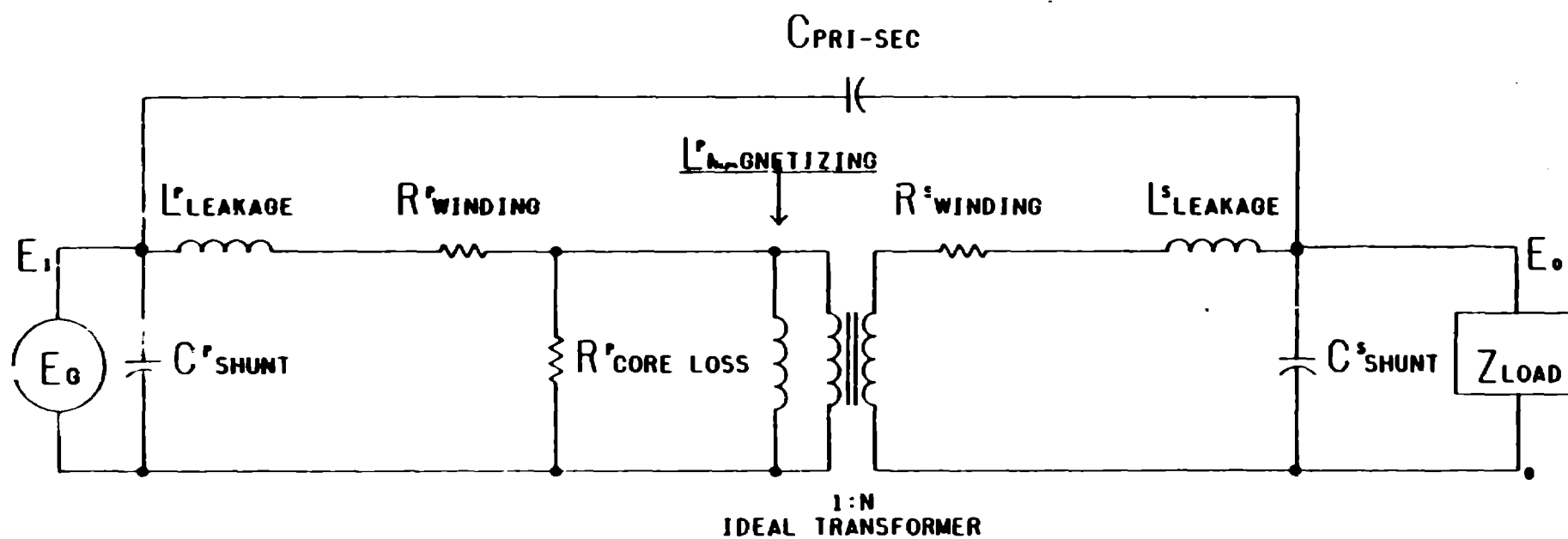


FIG. 21: GENERAL TRANSFORMER EQUIVALENT CIRCUIT

In this equivalent circuit in Fig. 21, all these lumped equivalent elements can be measured. For example, R_{gen} is the equivalent AC resistance in series with the generator; C_{shunt}^S is the shunt capacity of the secondary -- it's a measure of the lumped shunt capacity across the secondary, and it is ascertained in the same way as for the primary (i.e., by injecting a small current into the secondary and measuring the natural resonant frequency in resonance with the secondary leakage inductance with the primary shorted). Manufacturers of transformers for pulse applications can almost always give all these values to within 10-15%, which is generally satisfactory. If they cannot estimate them, then one should really consider whether he wants to buy from that source. One very important point to note is that there is no electrostatic shield shown in Fig. 21. Since this represents an ideal transformer, there is no worry about that. It is suggested to put electrostatic shields on the primaries of all transformers. There's no reason the manufacturer can't do it, but for separate primary and secondary bobbins, the individual can do it if he wishes. If the transformer primary is being referred to someplace other than ground, that electrostatic shield could quite legitimately be referred to that same place, recognizing that any current passing into it is going to traverse whatever other components are in that return loop to ground.

Now is the time to simplify this complex circuit. In the high voltage case, shielding is really intended to isolate the primary from fast transients that are being coupled back from the load, across the stray capacity of the filter inductor, back through the diodes (because they're shunted with capacitors), back through the transformer and into its primary. The advantage of the electrostatic shield on the transformer is that these transient currents travel into the shield and are harmlessly directed off to the ground of the transformer assembly. This shield then protects the primary from being exposed to such fast transients. Similarly, a fast transient coming into the primary side from the line cannot couple to the transformer secondary where it may become large enough to cause damage.

In the case of the ideal transformer, because there is conservation of power, the output current is the input current over N (Fig. 22). In the real world, referred to the primary, since dissipated power is conserved when it is reflected from the secondary loss elements, all losses can be expressed as equivalent circuit elements in the primary, as illustrated in Figs. 22 and 23.

If the primary current is i_1 and the secondary current is i_0 , the same equivalent primary loss for power $i_1^2 R'$ equivalent is sustained for both of these. So the overall equivalent resistance in the primary, taking into account the resistive loss in the secondary as well as the primary loss resistance gives:

$$R = R_{\text{winding}}^P + \frac{1}{N^2} R_{\text{winding}}^S$$

Whatever loss resistance is in the secondary can be referred back to the primary side, so lumped loss elements can be used for energy-dissipation calculations. This lumped equivalent model is a way of moving dissipative terms, capacitances, and inductances from one side to the other through ideal transformers to make calculations a little more straightforward. The power loss terms can be referred to the secondary or to the primary. This is an elementary way of referring the lumped elements to the primary side from a loss point of view, an equivalent shunt capacity point of view, and from an equivalent inductance point of view. To reflect the other elements back into the primary, refer to Fig. 22 for the equivalents. This circuit comes from Terman and there is a very good discussion in Fink's Radio Electronic Engineering.

When all these lumped elements are reflected to the primary at low frequencies, the circuit of Fig. 23 results. At low frequencies the capacitances are not required. R is the winding loss term for both primary and secondary losses and R_{core}^P loss is the core loss term. The total magnetizing inductance is $L_{\text{magnetizing}}^P$, the load is Z_{load}/N^2 , and this circuit equivalent is applicable from about 60 Hz to 5 kHz. It's a good circuit for the transformer and can be used for most power supply design.

FIG. 22: SIMPLIFIED TRANSFORMER EQUIVALENT CIRCUITS:
REFERRED-TO PRIMARY

IDEAL

$$E_1 I_1 = E_o I_o \quad \therefore \text{POWER CONSERVED}$$

$$E_o = N E_1$$

$$I_o = \frac{I_1}{N} \quad 1:N = \text{TURNS RATIO}$$

REAL: REFERRED TO PRIMARY

CONSERVING POWER DISSIPATED:

$$A. \quad I_1^2 R'_{\text{equivalent}} = I_o^2 R_{\text{winding}}^s$$

$$\therefore \underline{R'_{\text{equivalent}}} = \frac{1}{N^2} R_{\text{winding}}^s$$

$$B. \quad \frac{1}{2} E_1^2 C'_{\text{equivalent}} = \frac{1}{2} E_o^2 C_{\text{shunt}}^s$$

$$\therefore \underline{C'_{\text{equivalent}}} = N^2 C_{\text{shunt}}^s$$

$$C. \quad \frac{1}{2} I_1^2 L'_{\text{equivalent}} = \frac{1}{2} I_o^2 L_{\text{magnetizing}}^s$$

$$\therefore \underline{L'_{\text{equivalent}}} = \frac{1}{N^2} L_{\text{magnetizing}}^s$$

$$D. \quad \frac{E_1^2}{Z'_{\text{equiv}}} = \frac{E_o^2}{Z_{\text{load}}}$$

$$\therefore \underline{Z'_{\text{equiv}}} = \frac{1}{N^2} Z_{\text{load}}$$

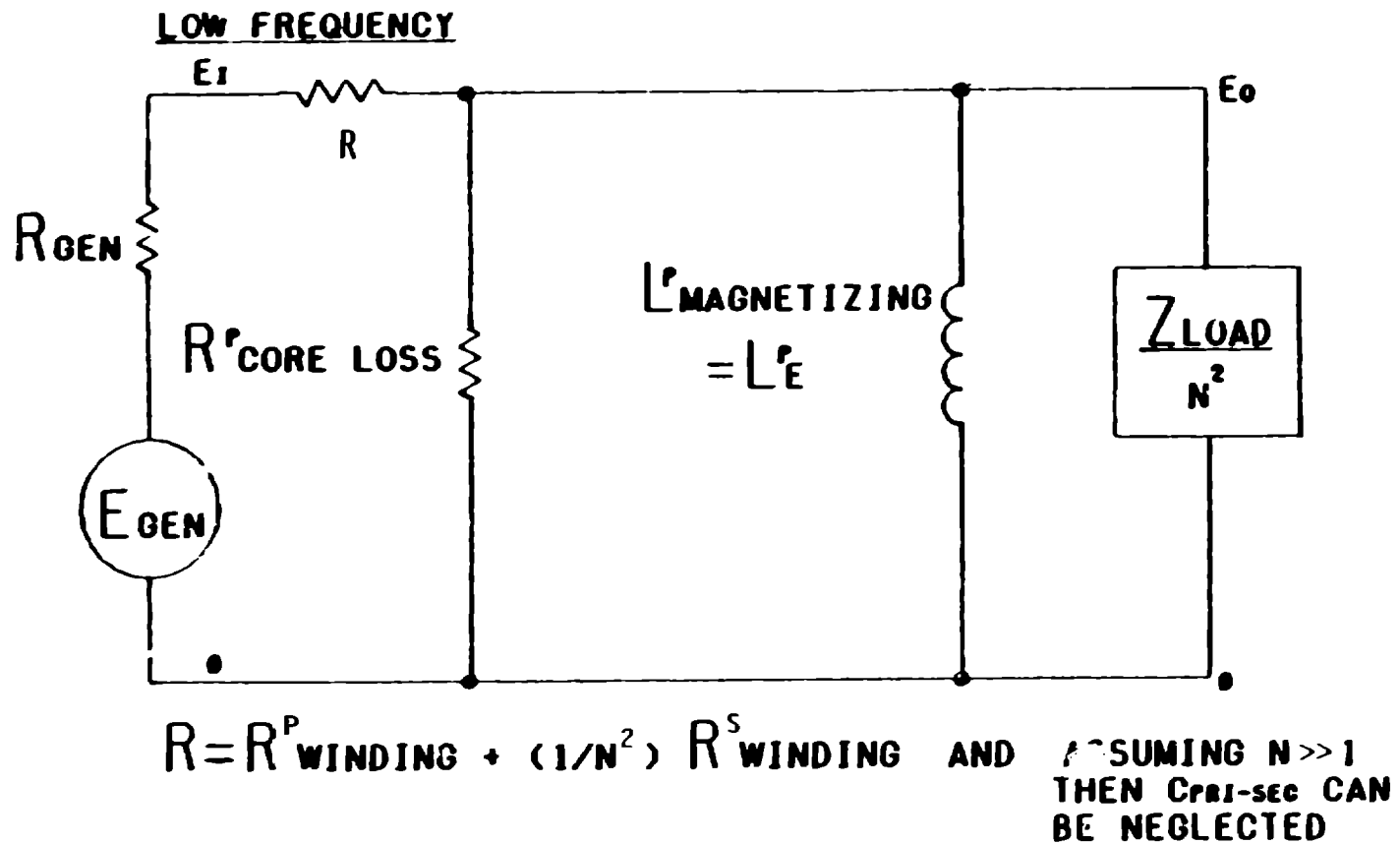


FIG. 23: SIMPLIFIED LOW-FREQUENCY TRANSFORMER EQUIVALENT CIRCUIT REFERRED TO THE PRIMARY.

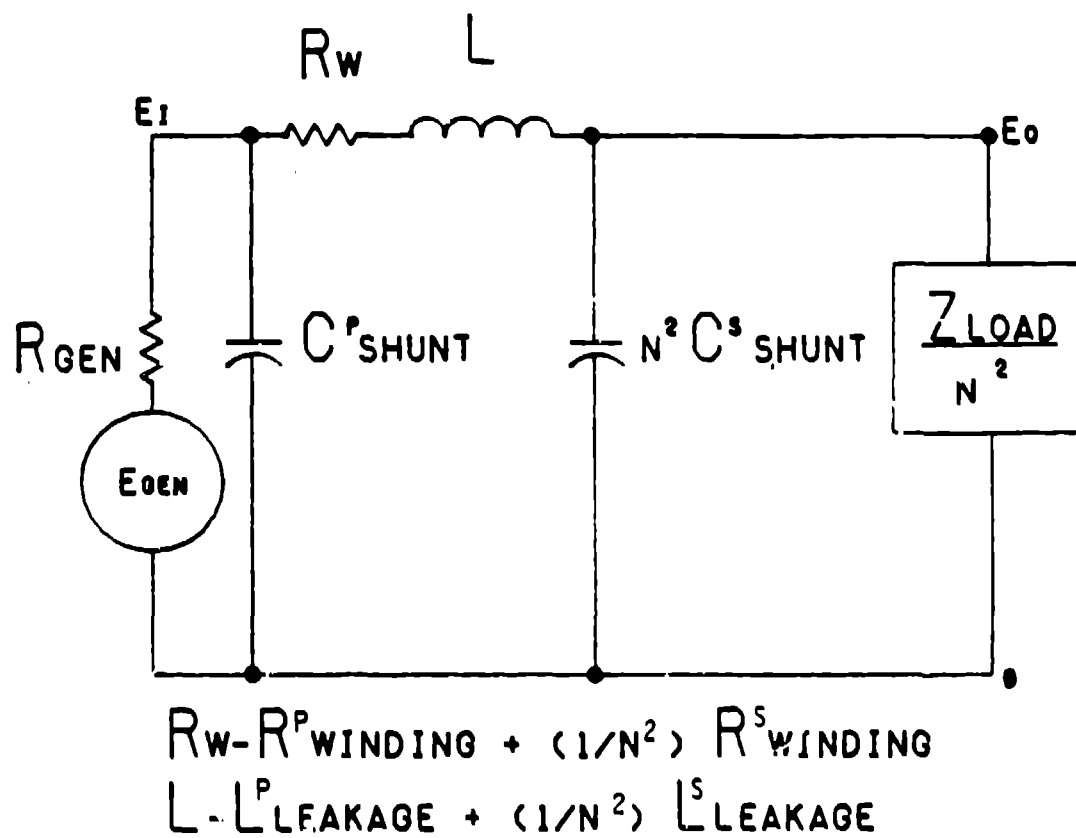
At high frequencies the second equivalent circuit (Fig. 24) is obtained. Shunt resistive and inductive components are no longer of concern, so further simplifications arise. The magnetizing inductance term was eliminated (the core loss still exists, but is relatively insignificant in terms of the energy stored in various shunt capacitances and leakage inductances). The L is the sum of all the leakage inductances. The winding loss and generator resistance remain although they may now be frequency dependent. The secondary shunt capacity is referred back to the right side of the equivalent leakage inductance as $N^2 C_{shunt}^s$. That's important as now we have the potential for transient excitation of this dampened series RLC oscillator. The idea now is to dampen these oscillations and try to keep them under control so that the voltage nowhere exceeds twice the peak applied voltage. This amplitude limit is selected primarily because it gives a good cost/performance trade off.

There is now a resonant circuit as illustrated in Figs. 25 and 26, where the the generator voltage is E_{GEN} . Solving the circuit equations at the resonant frequency $\omega^2 LC = 1$ and $Q = \omega L/R$. At resonance, the circuit is pure resistive, the current is a maximum, so the voltage across the resistor is E_{GEN} .

What about the voltage across the inductor? If this voltage is desired to be twice E_{GEN} , then:

$$V_L = j Q E_{GEN} = j 2 E_{GEN}$$

Now, for the capacitor, since $V_C = -j \cdot Q \cdot E_{GEN}$ or $= -j \cdot 2 \cdot E_{GEN}$ (for a $Q = 2$ circuit), the peak voltage across any element is never greater than twice the peak generator (or transient) voltage. That's all there is to snubbers. No matter what the initial conditions are, no matter what is done, there's no way for more voltage to appear across the inductor than twice the peak value of the AC voltage that is in the circuit. There is, however, one rare exception: there's a very special category of circuits called "two-frequency oscillatory discharge networks" and the one time they



NOTE: R_{GEN} AND R_W MAY BE FREQUENCY DEPENDENT.

FIG. 24: HIGH-FREQUENCY TRANSFORMER EQUIVALENT CIRCUIT

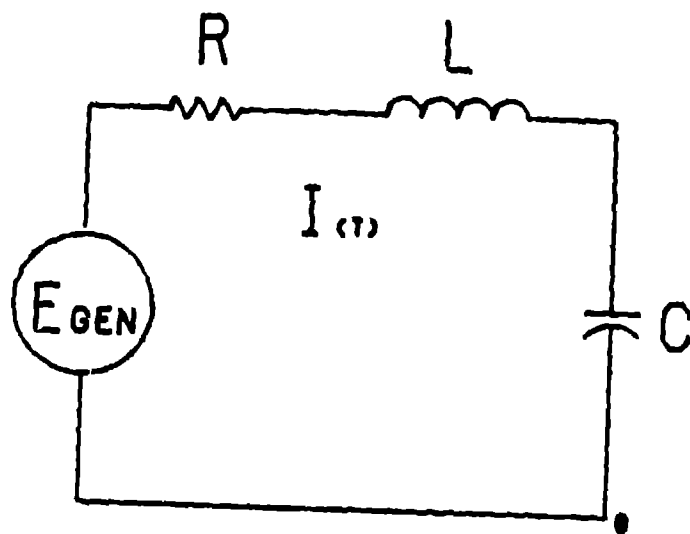


FIG. 25: THE SERIES RESONANT CIRCUIT

FIG. 26

$$E_{\text{GEN}} = E_0 e^{j\omega t}$$

$$\text{SET } I = I_0 e^{j(\omega t + \phi)}$$

THEN, AT TIME T FOR $I = 0$ AND NO "Q" ON C:

$$\begin{aligned} E_{\text{GEN}} &= I R + L \frac{dI}{dt} + \frac{1}{C} \int I dt \\ &= I_0 e^{j(\omega t + \phi)} R + L I_0 e^{j(\omega t + \phi)} j\omega + \\ &\quad \frac{1}{C} \left[I_0 \int e^{j(\omega t + \phi)} dt \right] \\ \therefore \int e^{j(\omega t + \phi)} dt &= \frac{1}{j\omega} e^{j(\omega t + \phi)} \Big| = \frac{1}{j\omega} \left[e^{j(\omega t + \phi)} - e^{j\phi} \right] \\ \therefore E_{\text{GEN}} &= I R + j\omega L I + \frac{1}{j\omega C} \left\{ I - e^{j\phi} \right\} \end{aligned}$$

SINCE AT $T = 0$, $Q = 0 \therefore \phi = 0$:

$$\therefore E_{\text{GEN}} = Z I \text{ WHERE } Z = \left\{ R + j\left(\omega L - \frac{1}{\omega C}\right) \right\}$$

$$\text{SET } \omega^2 L C = 1 \longrightarrow Z = R \longrightarrow E_{\text{GEN}} = R I$$

\therefore CIRCUIT PURE RESISTIVE

$$\text{NOW } V_L = j\omega L I = j\omega L \frac{E_{\text{GEN}}}{R} = j\frac{\omega L}{R} E_{\text{GEN}}$$

$$\text{DEFINE } Q = \frac{\omega L}{R} \therefore V_L = Q E_{\text{GEN}}$$

I.E. V_L ACROSS THE INDUCTOR
LEADS E_{GEN} BY 90 DEGREES AND
IS "Q" TIMES LARGER.

FIG. 26A

NOW LET US RESTRICT THE PEAK VALUE OF THE VOLTAGE ACROSS THE INDUCTANCE TO $2E_0$.

$$\text{I.E., } (V_L)_{\text{PEAK}} = j Q E_0 = j 2 E_0 \rightarrow Q = 2$$

IN THIS CASE, THEN, $\omega L = 2R$

AND ACROSS CAPACITOR:

$$V_C = \frac{1}{j\omega C} I = \frac{1}{j\omega C} \cdot \frac{E_{\text{GEN}}}{R} = -j \frac{\omega L}{R} \cdot E_{\text{GEN}}$$

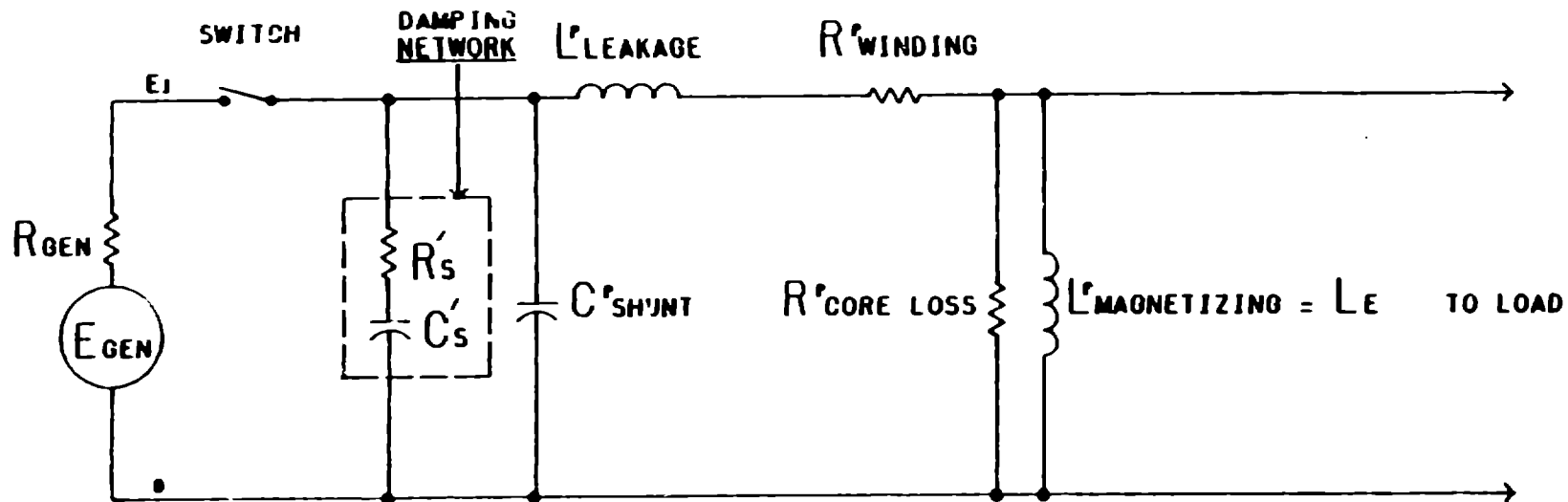
$$V_C = -jQ \cdot E_{\text{GEN}}$$

THUS THE PEAK VALUE HERE IS ALSO $2E_0$.

will be a problem is in the case of ferroresonances in power transformers. With most of the problem areas of interest in this course, there'll never be a worry about this. It's a problem that comes about when circuit breakers don't always open simultaneously or they don't close simultaneously, or when people insist upon building power supplies that are three phase and do not put phase drop-out detectors in them. When ferroresonance like that occurs, enormous voltages can be generated in the primary of the transformer. That's the only time not to believe the above analysis. Greenwood has an excellent discussion of this topic. To avoid that, just ask every power-supply supplier who's supplying three-phase supplies to put in a phase-drop or voltage detector. Then the problem goes away.

The damping network to be used for primary damping is shown in Fig. 27. The leakage inductance, L_E , is measured this way; $(E_1)_{rms}$ is applied and $(I_1)_{rms}$ measured for the secondary open circuit. From the previous analysis it can be shown that L_E can be calculated (Fig. 19). Just at turn-on there is no load on the power supply and the main inductance of the circuit is the magnetizing inductance. A series resonant circuit is then being excited, there is a shunt capacity C_{shunt}^P in series with the leakage and magnetizing inductances. The resonance can be dampened by inserting a shunt-loss term across the C_{shunt}^P . The resonant frequency of the circuit is 60 Hz. Now from $Q = 2$, the value of the shunt capacitance C_S' can be derived (Fig. 27) from $\omega^2 L_E C_S' = 1$ (generally $L_E \gg L_{leakage}^P$). The core loss, $R_{core\ loss}^P$ is known, so from Fig. 27 the series damping resistance R_S' can be calculated. It turns out that C_S' is generally about 100 times bigger than C_{shunt}^P . So there is how a resonant network here, $C_S' - L_E - (R_{core\ loss}^P + R_S')$ resonating at 60 Hz, dampened with a Q of two so it cannot oscillate with a higher Q at the natural resonant frequency of $C_{shunt}^P - L_E$, which could be several tens of kilohertz. Capacitors in the primary side, for C_S' are quite inexpensive. SCR snubber capacitors are generally used here because, although they cost about 20% more, they're of extended foil or multi-tab construction and their design lifetime is an order of magnitude longer than the less

FIG. 27: PRIMARY DAMPING FOR A REAL TRANSFORMER



1. DETERMINE $L_E = \frac{(E_i)_{rms}}{\omega(I_i)_{rms}}$ FOR NO SECONDARY LOAD. NOW $L_E \gg L_{leakage}^P$
2. RESONATE C'_S WITH $L_E \Rightarrow \omega^2 L_E C'_S = 1$ $F = 60 \text{ Hz}$
 FOR $V_{peak} = 2 (E_i)_{peak}$ $Q = 2 \therefore \frac{\omega L^P}{R^P + R'_S} = 2$ WHERE $R^P = R_{winding}^P + R_{core loss}^P$

SINCE $R_{core loss}^P$ IS KNOWN, THEN R'_S CAN BE CALCULATED.

expensive ones. An SCR (Silicon Controlled Rectifier) snubber capacitor is a pulse discharge capacitor made by the thousands. The people who make cycloinverters, for example, would be in serious difficulty if one of those failed. If one fails, this can destroy the whole SCR stack very quickly. Capacitor manufacturers will sell them at a modest premium over the cheaper filter capacitor design.

The secondary, depending upon the conduction angle of the diodes, for a very short time has a chance of being unloaded as the supply is turned on. That's one way of generating secondary transients, so a secondary snubber across the transformer is highly recommended. The same $Q=2$ constraint will be applied to the secondary damping case. The secondary damping is described in Fig. 28. Fig. 29 shows the equivalent circuit and an analysis of damping characteristics by frequency shifting.

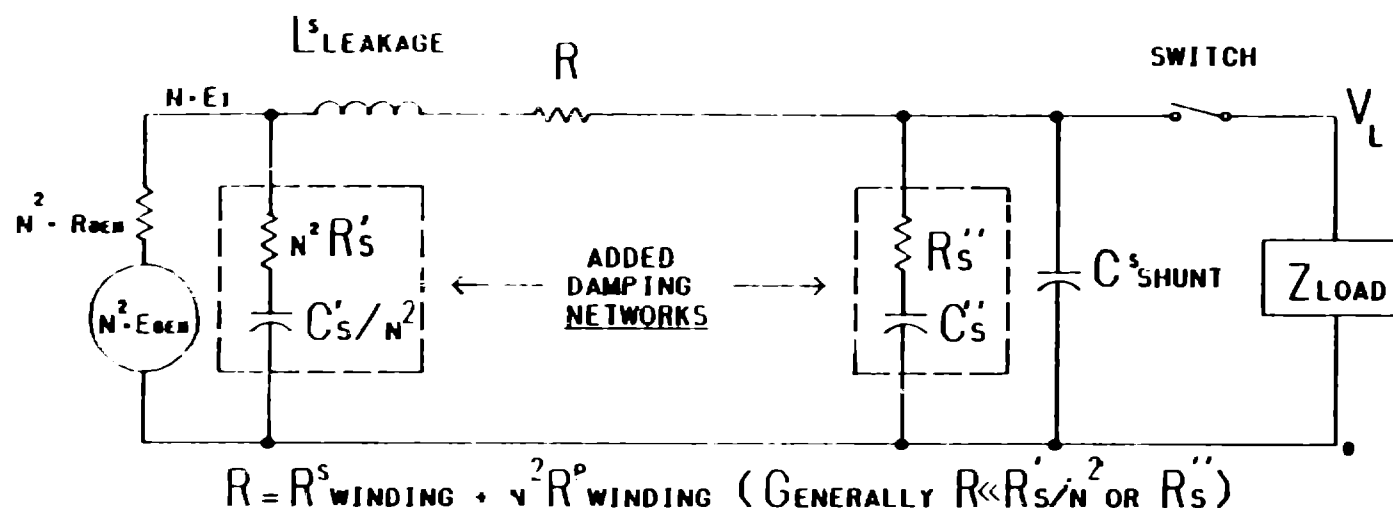
If the line voltage is turned on faster than the diode turns on (and there is a good possibility of that happening) -- the coupling from primary to secondary via $L_{leakage}^S$ and C_{ps} tend to do that -- oscillations in the secondary circuit can be shock excited at the resonant frequency of the transformer secondary with its shunt capacity. As an observation, experience has generally been that damping the secondary is far more important than damping the primary. However, it's more expensive. The difficulty is with the physical size of the capacitors needed in relatively high-voltage power supplies. For example, consider a 250-kw, 100-kV transformer system, that would probably be a six-phase Y-delta stacked configuration. (There is a / secondary and FWB and delta secondary and FWB and the two are added in series.) In that type of circuit C_s' of values up to 50 nanofarads for the snubber are common. This now represents a significant reactive power flow through the snubbers and, at these voltages, the damping networks become quite expensive.

The secondary resonant frequencies tend to be in the 20 to 100-kHz range. The resonance is between the leakage inductance and C_{shunt}^S , with resistive damping from $R_{winding}^S + R_{core}^S$ loss. There can be a large peak voltage -- Q times the peak voltage in steady state at turn-on. Snubbers should be added to control these voltages and then

FIG. 28: SECONDARY DAMPING

THE APPLICATION OF PEAK LINE VOLTAGE AT A RATE FASTER THAN DIODE TURN-ON (I.E. SUPPLY IS OPEN CIRCUITED) CAN GENERATE AN OSCILLATORY DISCHARGE IN THE SECONDARY CIRCUIT, ESSENTIALLY AT THE NATURAL RESONANT FREQUENCY OF THE TRANSFORMER (FROM 20,000 TO 100,000 Hz). THE FREQUENCY DEPENDS UPON THE SECONDARY "LEAKAGE" INDUCTANCE IN SERIES WITH INTERNAL SHUNT CAPACITANCE. THE CIRCUIT IS OF A HIGH Q AND CAN GENERATE VERY HIGH PEAK VOLTAGES ($\approx Q \cdot V_{SEC}^{PEAK}$). THESE TRANSIENT PEAKS CAN BE REDUCED TO REASONABLE AND CONTROLLABLE DESIGN LIMITS, WITHIN INSULATION RATINGS OF TRANSFORMERS (AND DIODES) TO, SAY, $2 V_{SEC}^{PEAK}$, BY ADDING AN RC SNUBBER ACROSS THE SECONDARY OF THE TRANSFORMER. THE VALUES ARE OBTAINED FROM THE FOLLOWING MODEL, DERIVED FROM THE SERIES RESONANCE CIRCUIT THAT WAS BRIEFLY DISCUSSED BEFORE.

FIG. 29: HIGH FREQUENCY SECONDARY DAMPING



1. NATURAL SELF RESONANT FREQUENCY: $\omega_R L^S C_T = 1$ WHERE $C'_S = 0$; $C_T^{-1} = (C^S)^{-1} + \frac{C'^S}{N^2}$
2. NOW: FOR A PEAK OVERVOLTAGE OF 2x WHEN SWITCH OPENED THEN SHIFT THE RESONANCE BY N TIMES IN FREQUENCY (5x ABSOLUTE MINIMUM). IN THIS CASE IT CAN BE SHOWN THAT THE PEAK VOLTAGE AT ω_R REMAINS BELOW $(N E_i)$ =

$$\left\{ \frac{1}{\frac{1}{C^S_{\text{SHUNT}} + C'^S_S} + \frac{N^2}{C'^S_S}} \right\} = C^S_{\text{SHUNT}} = 100 \quad \text{ALSO, } 2 R_T = \sqrt{\frac{L^S}{C_{T0}}}; \therefore Q = \frac{\omega_0 L^S}{R_T} = 2; \omega_0^2 L^S C_{T0} = 1$$

SHUNT

$$R_T = R'^S_S + N^2 R'_S + R, \text{ WHICH DEFINES } R'^S_S.$$

values are derived from the model of the series resonant circuit, which was briefly discussed before (Fig. 25).

This is interesting because this particular general circuit, with its damping snubbers, appears everywhere throughout PCS design. Snubbers are used on some laser systems, magnetrons (often for mode control), or klystrons, and it's always the same argument. What we're going to try to do here is come up with a general way of deriving the values for C'_S and R'_S (C'_S and R'_S having been determined in the preceding (Fig. 30) analysis). There are a lot of cookbook prescriptions for these snubbers that are found in the various diode manuals or the transformer literature, but they tend to overcompensation techniques, using large capacitors and modest values of resistors. This technique here allows considerable economy, because capacity in high-voltage capacitors costs money. If there's any significant RMS current (more than an amp or two) it does cost a lot.

It can be shown that, starting from the $Q(\omega)$ analysis in Terman, if a snubber network is designed whose resonant frequency is 1/10 the natural resonant frequency, there is no way that for any Fourier components of any pulse of any shape that is applied to the circuit to generate a voltage across any component of more than twice the peak voltage of that Fourier component. The circuit lumped equivalent elements are now all referred to the secondary to simplify the circuit in Fig. 29. We've referred the primary damping capacitor, the primary damping resistor into the secondary. $L_{leakage}^S$ is the leakage inductance, R is the total winding loss, C_{shunt}^S is the shunt capacitance of the transformer. There's also an output switch: it could in fact be a diode. The power supply is turned on, the diode closes, and it conducts current into the load. The transformer load is in this case the filter inductor, the filter capacitor, and the load resistor that is attached across the supply.

The natural resonant frequency occurs when this $L_{leakage}^S$ is in resonance with the C_T , which is $(C_{shunt}^S + C'_S)$ in series with C'_S/N^2 .

Initially, the resonant frequency ω_R is calculated for zero snubber (C_s'') capacitance. The analysis is shown in Fig. 29. For the peak overvoltage of twice the maximum secondary voltage when the switch is opened, we proceed as before and shift the resonant frequency downwards by a factor of ten. Then the peak voltage at ω_0 , the new resonant frequency, always remains below the peak secondary voltage. Shifting ω_R by only a factor of five means that the peak voltage at this new resonant frequency ω_0 could be twice the peak secondary voltage. For a shift in resonant frequency of ten times, C_s'' and R_s'' are determined from the relations in Fig. 29. If the resonant frequency in the secondary is shifted by ten times, then it can be shown that nowhere can there be more peak volts than the peak volts from this new resonant frequency, even though as higher frequencies are reached higher Qs appear. Terman provides a lucid proof of this. This is important because it means no matter what happens there's no way the total peak voltage (the sum of these transformer and transient peak voltages) can then be more than twice the peak secondary voltage across any component. There are different self-resonant frequencies in the primary and the secondary of the transformer and normally they differ by a significant amount. They are determined primarily by the shunt capacities and the leakage inductances.

Everything we've discussed so far can also be used for designing pulse transformers. The same arguments apply for snubbing and control of oscillations in pulse transformers. Rise and fall-time aspects will be considered later. If the resonant frequency is shifted by ten times, and if Q is equalled to 2, there is a new ω_0 . C_{T0} is known, then if that is divided by C_{shunt}^s and equated to 100 (this ensures that the resonant frequency is shifted by 10 times), then the resonances are going to be well-behaved and there won't be more than twice the peak applied volts. For $Q = 2$, then:

$$2R_T = \sqrt{\frac{L_s}{C_{T0}}}$$

With R_T known, then R_s'' can be calculated. C_{T0} provides C_s'' and the snubber values are now known. Typical values are kilohms

and fractions of a nanofarad. Analysis reveals that the added KVAR loading from the snubber is around 5% for $Q = 2$.

If it is desired to eliminate the snubber capacitors entirely, they can be replaced with nonlinear elements called thyrites or metal-oxide varistors (MOV). Basically, what these are are non-linear resistors that have a characteristic where above a certain critical voltage the current increases rapidly. They're essentially similar to symmetrical Zener diodes, with a finite lifetime as a percentage of pulse overloads. The lifetime shortens drastically at high-peak currents, making it necessary to control the peak fault currents through them very carefully in order to obtain reliable, long-life operation. The cheapest way to do this and the one that works very well is to put them across the snubber capacitors so you can reduce their capacitance value. Even if more voltage appears across that capacitor (as the Q is higher then) the varistors protect it by clamping the voltage level. They are difficult to use in large power supplies. They can only dissipate a few watts (up to 50 W), have significant capacitance, and are quite expensive. At higher voltages they're as expensive as the snubber capacitors. The normal MOV acts like a 0.01 to 0.1 microfarad capacitor, so there is a free damping capacitor. One type that is rather useful is a zero-capacity MOV. Zero-capacity MOVs are valuable for pulse sharpening applications and have nanosecond risetimes. They're made by a different process and have a 6th power law increase in current with voltage above the knee value, at a very low capacity, making them extremely useful for shunting transients and RFI in circuits when there are large peak-to-peak voltage problems at high frequencies. They can carry a fair amount of RMS current at high frequencies.

This, then, is how the transformer secondary can be dampened. The capacitor C_s resonates with the series RLC circuit at $1/10$ the natural resonant frequency, and the resistor R_s is selected so the Q is 2 (arbitrary choice to control the voltage to within a factor of 2 everywhere). The same argument applies to the filter choke damping.

Typical values for the R_s'' of the right side are kilohms and typical values for the C_s'' are fractions of a nanofarad. For interest sake, at the 100 kV level a 1-nF capacitor is a value above which costs escalate rapidly. If 10 nF is desired, the price goes up very quickly. There's clearly a tradeoff required. Fig. 18 illustrates the effect of leakage inductance in generating voltage transients in the secondary circuit.

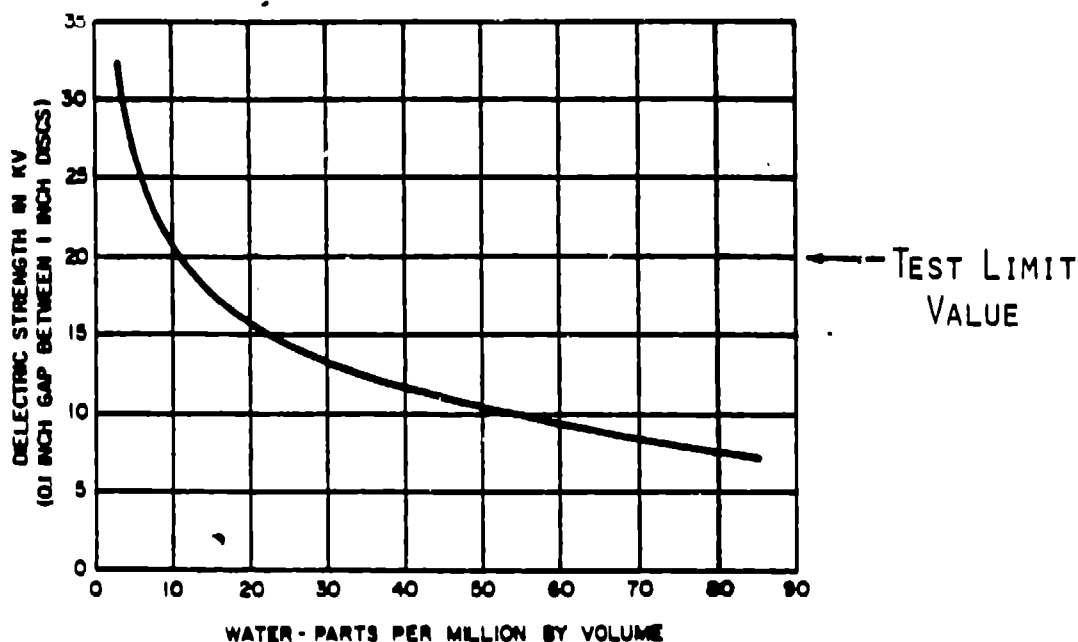
For 60 Hz to 360 Hz, kilovolt power supplies, thyrites and MOV are quite useful devices. They can be used to build diode stacks in disaster situations to replace the capacitor/resistor combination in the compensative diode stacks just discussed. For 1 kV diodes, the 750 V standard MOV units can be used although they are rather expensive.

OIL INSULATION

The Fig. 30 is a simplification of much of the insulation data in Mlynar's book. A study of these points may save a great deal of agony and lost time. Note that if there are even a few parts per million of water in oil and the water stays suspended, the dielectric strength in kV per 0.1 inch drops drastically. People who make single-shot generators (e.g., large Marx banks) really don't have to worry about that. When these large devices are assembled the water sinks to the bottom. When dealing with small-size assemblies, particularly at high rep-rates, the AC corona component is significantly enhanced by the presence of water in the oil. Anyone contemplating any high-voltage work is advised to acquire a small oil purifier. If at all possible it is advisable to purify the oil whenever it is taken out. It's worth the trouble and the small investment in time. The typical oil sitting around the lab or outside, even in "sealed" containers rapidly (\approx 6 months) can become contaminated to such an extent that the dielectric strength is seriously degraded. The oil purifier is a small centrifuge device that separates impurities from the oil, possibly followed by an addition of a particulate and water filter.

Antioxidants put into high voltage transformer oils react readily with a large number of components of which high-voltage assemblies are made, such as resistor varnishes and the wax coatings on cheap

FIG. 30: OIL INSULATION DATA



1. DESIGN FOR 30,000 VOLTS PEAK PER INCH.
2. CLOSER SPACINGS REQUIRE INSULATION BARRIERS BETWEEN HIGH-VOLTAGE POINT AND GROUND.
3. OIL EXPANSION WITH TEMPERATURE IS SIGNIFICANT.
4. FOREIGN MATTER CAN CONTAMINATE THE OIL (E.G. WAX-COVERED COMPONENTS, DIRT, PAINT, ETC.)
 - CAN REDUCE DIELECTRIC STRENGTH (AC, DC, OR PULSE)
 - CAN INCREASE LOSSES AT HIGH FREQUENCIES (E.G. RF CIRCUITS)
5. DISPLACE AIR IN TANKS WITH DRY NITROGEN OR ARGON. FREONS CAN BE USED IN EXPERIMENTAL SET-UPS (BUT FAR, FAR AWAY FROM CO₂ LASERS.)
6. OVERSTRESSES CAUSE WHIRLPOOLING AND LEAD TO ARCS.

capacitors. These antioxidants and other materials dissolved in the oil can polymerize, allow tracking, and carbonize insulating surfaces. If components are to be used inside high-voltage assemblies, wax-covered components are not recommended unless specifically designed to work in oil (the manufacturer provides this information). There are components designed for use in the standard high-voltage oils.

The recommended stress level near 60 Hz is 30,000 volts peak per inch. At high frequencies, this decreases by a factor of 10 (at a few megahertz). With Shell Dialax, above a megahertz in RF systems becomes a problem. In dealing with high-rep-rate circuits with very high-frequency components in them, consider using the new silicone liquid dielectric insulating fluids from Dow Corning: DC 200, which has been available for years, and a new one, DC 501, that is about a factor of four more expensive than Shell Dialax and is quite impressive. It has a low corona inception point and a very low loss factor at high frequencies. Some people are sensitive to oils, so if you get them on your hands, it's recommended to wash often with pumice soap.

The antioxidants in high-voltage oils like Shell Dialax can degrade over a period of years under relatively high electric field stresses. Therefore, old oil should be rejuvenated with an additive before it is used. Over long periods of time, the hydrocarbons cross-link, combining to lose their ability to function as a high-quality insulator. They all depend upon molecular structure to absorb the gas molecules from corona areas and this structure changes with time. If the oil is filtered and the antioxidants replenished, the oil can be rejuvenated. Oil should not be heated above 120° F. If it is, and then new transformers are impregnated in it, it is changed into something else that isn't as good (its long-term antioxidation properties, its long-term breakdown, and its long-term corona resistance are all poor) and the system can then fault. Silicone fluid can be heated (although there's no need to) to 240° F without damage; there's no damage until very high temperatures are reached. Mineral-based oils are also very stable and can be highly

abused. The best oils for lab applications are Dow Corning 200 and 561. Given a resistor about two inches long that is normally rated for 10 kV dc, when it was put in Dow Corning 200, the resistor sustained 110 kV dc indefinitely. With Shell Dialax the resistor tracked at about 30 kV. Dow Corning 200 is very useful material for lab applications, but may be too expensive for commercial applications. There are some insulating dielectrics better than the silicone fluids. They're very expensive and for a given geometry they can sustain a very high DC voltage before any significant corona occurs. That's what's really good about them. Their loss is generally high at high frequencies. The real advantage of silicone fluids is that they are very low loss up to 200-300 megahertz, and even then there's a transition to only a slightly higher loss. Silicone fluids (e.g., diffusion pump oils) can be rerefined by the manufacturer and will work well afterwards.

Closer spacings and higher stresses than the above require insulation barriers between the high voltage point and ground. If this is to be done, be sure that the insulating spacers do not suffer corona degradation or polymerize. One of the best materials known for this use, with high mechanical strength, is made by Permali. The possible exception to Permali might be in multimegahertz generators in which there are a large RF voltages. Then it might tend to break down and Teflon or Kapton may be preferable.

Oil, unfortunately, expands with temperature. This is obvious, but I once made a thyratron driver module of 1/2-inch thick brass in a fully-sealed geometry. The leads blew off the top. Silicones are considerably better in this respect than standard transformer fluids.

Garbage can dilute the dielectric strength of oils. Rosin in solders can polymerize and cause tracking. If possible, clean soldered items before putting them into oil (e.g. in Chlorothene). Rosin solder is a very bad material. The nonactivated flux that Kester and others sell for soldering printed circuit boards, a very low viscosity material, is very

good in terms of DC leakage and corona damage resistance. The standard rosin-core solder, particularly the activated types are to be avoided if fairly high voltage stresses abound.

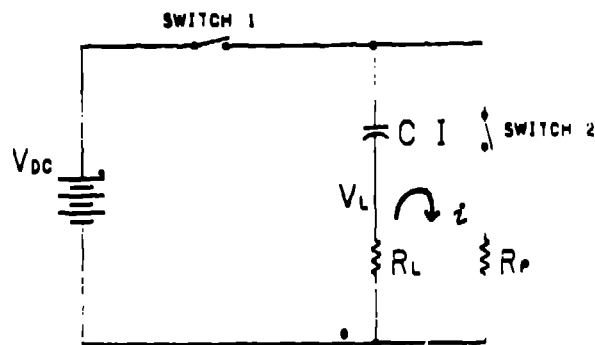
To minimize water vapor from the air passing into insulating fluids, the air in the high-voltage assembly is often displaced with dry nitrogen or argon (or Freon if there's no CO₂ laser around). Freon-12 dramatically suppresses surface tracking, keeps the water vapor out, displaces all the air, and won't allow air into the transformer.

Electrical overstressing causes whirlpooling in insulating fluids. Above certain stress points, the oil forms a whirlpool. For the lab application it generally doesn't matter, but this level of stress will polymerize the oil and cause degradation in hold-off capability and is not a recommended approach for long-life systems. This whirlpooling is caused by electron or ion field emission from sharp edges where the ions push the molecules into a whirlpool. The high emission currents and fields also polymerize the oil. The whirlpool is simply an indication of other factors that are not obvious but are occurring simultaneously. A typical number for time limits in a test system at 100 kV was days. After weeks of operation deposits occurred causing numerous arcs. The other thing whirlpooling does is pushes the oil up the container surface and out onto the floor.

III. HARD TUBE POWER CONDITIONING SYSTEMS

A hard tube pulser can be represented as an analog switch in series with a battery. To a very good approximation, (Fig. 31) the capacitive storage hard-tube pulser is a large reservoir capacitor with a series switch, some stray capacitance across the load, an isolating inductance for recharging, and some type of load. The inductive storage type is illustrated in Fig. 32. The balance of this discussion talks about what happens when the switch (1) in Fig. 31 is opened and how to

FIG. 31: HARD TUBE PULSERS



1. CLOSE SWITCH 1 TO CHARGE C;
SWITCH 2 OPEN:

$$V_C(t) = V_{DC} (1 - e^{-t/R_L C}) - V_C(0)$$

2. OPEN SWITCH 1 AND CLOSE SWITCH 2:

$$V_L = V_{DC} e^{-t/RC} - V_P$$

SINCE $R = R_L + R_p$ AND $V_P = I_P R_p$, SWITCH VOLTAGE DROP = CONSTANT

NOW, LET SWITCH 2 BE CLOSED FOR TIME $\tau \ll RC$. $\Delta V_L = \frac{V_{DC}}{RC} \cdot \tau$

DURING WHICH TIME SWITCH 2 DISSIPATES POWER.

FOR THE TUBE SWITCHED FULLY ON, THE TUBE DROP IS V_P AT PEAK CURRENT I_P . THE ENERGY LOSS PER PULSE IS $V_P \cdot I_P \cdot \tau$. AVERAGE POWER LOST PER PULSE IS THEN $P_{AV} = W_P \cdot PRF$.

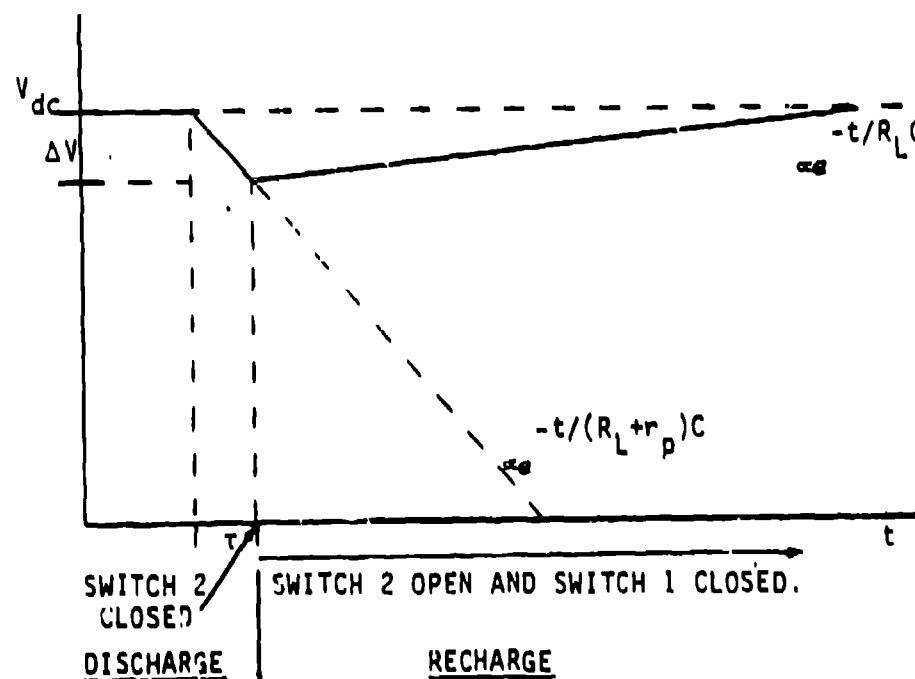
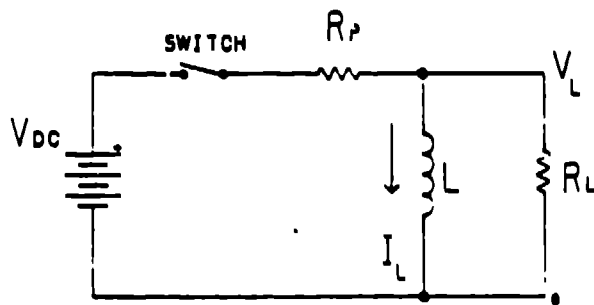


FIG. 32: POWER CONDITIONING SYSTEMS USING
INDUCTIVE ENERGY STORAGE



CLOSE THE SWITCH, THEN:

$$I_L(t) = \frac{V_{DC}}{R_P} \left(1 - e^{-\frac{R_P}{L}t} \right)$$

ASSUMING $R_P \ll R_L$ AND ZERO INDUCTOR RESISTANCE, AND OPENING THE SWITCH:

$$I_L(t) = \frac{V_{DC}}{R_P} \cdot e^{-\frac{R_L}{L}t} \quad (\text{ASSUMING THE SWITCH WAS OPENED AT PEAK CURRENT}).$$

THE PEAK LOAD VOLTAGE, $V_L = \frac{V_{DC}}{R_P} \cdot R_L$, WHICH CAN BE VERY LARGE FOR $R_P \ll R_L$ AND THE SWITCH MUST HOLD OFF THIS VOLTAGE AT THE INSTANT OF OPENING.

$$\text{I.E. } V_L = -L \frac{di}{dt} = - \frac{V_{DC}}{R_P} \cdot R_L \cdot e^{-\frac{R_L}{L}t}. \quad \text{AT } t = 0,$$

$$V_L = -V_{DC} \left(\frac{R_L}{R_P} \right)$$

NOTE THAT THE SWITCH CURRENT REMAINS ON DURING THE CHARGE TIME SO THE POWER DISSIPATION IS HIGH.

keep switch (2) dissipation at a minimum. Typical switch resistance is several hundred ohms. That means a low-impedance Blumlein can't be driven with a hard-tube pulser, so that hard tube pulsers are very good to drive a high-impedance load. In the kilovolt range, small planar triode tubes achieve only 50 Ohms when saturated, and even the kiloamp tubes that cost thousands of dollars are 100 Ohms or so. That's basically the space-charge limited current the cathode can produce.

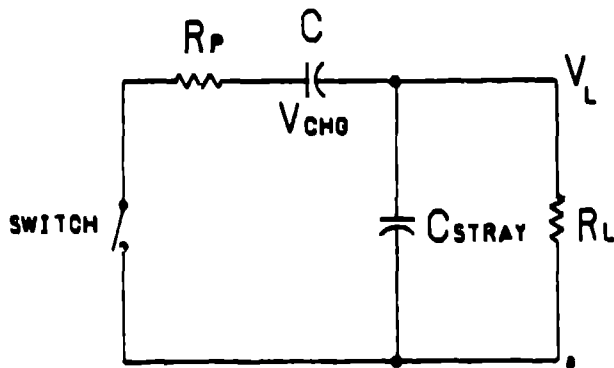
There are two switches in Fig. 31. If switch 1 is closed and C is charged up, the RC charging time of that is limited by R_L . Opening switch 1 and closing switch 2, then the series resistance in the circuit is $R_L + R_p$. If switch 2 is closed for a short time, τ , the change in the voltage across R_L is $V_{DC}\tau/RC$. The voltage is almost V_{DC} , but charge is being pulled out of this capacitor in an RC discharge. That's the prime switching cycle in a hard tube pulser.

When switch 1 or 2 is closed it is dissipating power. The power lost can be calculated. Given a voltage drop across the tube, V_p at the peak current I_p , then the energy lost in the switch per pulse is approximately $V_p I_p$. This times the PRF gives the average power dissipated, which can be a very large number. In a megawatt modulator, many kilowatts of average power can easily be lost into the switch.

Risetime is determined by the RC_{stray} time constant (Fig. 33) where R is a total equivalent resistance in series with C_{stray} , giving a risetime of that R times C_{stray} . Since generally for high-efficiency $R_p \ll R_L$, then $R \ll R_p$, which is the switch resistance. The high-power tubes are very large and have inductances of many microHenries. Risetimes for hard tube pulsers for such systems are several microseconds. This inductance does enter when the switch is opened in some circuits, sometimes giving rise to oscillations. If there is a switch with RC snubbers across it (added because it's misbehaving itself), there can be a RLC loop, in which the stored energy in that gives rise to large oscillations, causing the tube to arc and fault the DC supply. On

FIG. 33: RISETIMES AND FALLTIMES IN A HARD-TUBE PCS

70



A) $V_{\text{LEADING EDGE}}(\tau) = \frac{V_{\text{CHG}}}{R_P} \cdot R \cdot (1 - e^{-\frac{\tau}{RC_S}})$

WHERE $\frac{1}{R} = \frac{1}{R_P} + \frac{1}{R_L}$ RISETIME = $RC_{\text{STRAY}} = \tau_r$

FOR $R_P \ll R_L$ THEN $\tau \approx R_P \cdot C_{\text{STRAY}}$

FOR EXAMPLE, IF $R_P = 100 \Omega$ $C_{\text{STRAY}} = 20 \text{ PF}$.

THEN $\tau_r = 2 \text{ NS}$.

B) $V_{\text{FALLING EDGE}}(\tau) = V(\tau) e^{-\frac{(t-\tau)}{R_L \cdot C_S}}$

SINCE $R_L \gg R_P$, THEN $\tau_f = R_L C_S \gg \tau_r$ FOR EXAMPLE IF $R_L = 1000 \Omega$

THEN $\tau_f = 20 \text{ NS}$ IF $C_{\text{STRAY}} = 20 \text{ PF}$.

NOTE: IF $\tau > 3 \cdot R \cdot C_{\text{STRAY}}$, THEN $V(\tau) \approx \frac{V_{\text{CHG}}}{R_P} \cdot R$. ASSUMING THAT

THERE IS SMALL VOLTAGE DROP ON "C".

NOTE: WHEN CONNECTED TO A CHARGING SUPPLY AT POINT A THROUGH A RESISTOR R_{CHG} , THEN ON CLOSURE, IF $R_{\text{CHG}} \gg R_P$, IT HAS NO EFFECT ON τ_r . NOW, WHEN THE SWITCH IS OPENED, R_{CHG} IS IN PARALLEL WITH R_L AND CAN BE USED TO REDUCE THE FALLTIME.

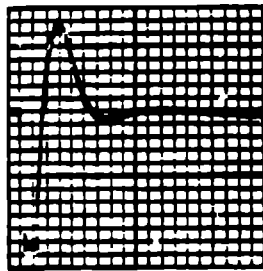
the falling edge, the switch is opened up and R_L is parallel with C_{stray} , revealing the falling-time constant.

When the power supply is added there is another resistive term to take into account. In inductive charging (Fig. 34) the inductor is placed across the load in series with switch 2, forming the DC return. The inductor is added to reduce the recharging losses and the output pulse appears on the load when the voltage exceeds the switch 2 closure voltage. The second switch (2) is closed (switch 2 could represent a laser load, for example, that stays on while there are so many volts across it and if the voltage on the load falls below the holding voltage suddenly switch 2 opens and the laser turns itself off. Magnetrons also behave that way.) When switch 2 opens there is another resonant circuit. If the Q of that is controlled correctly, there is no problem. If the Q is not low enough, a diode can be inserted across C_s as a shunt diode, clamping the inverse voltage, as shown in Fig. 34. The conditions in here can be chosen for critical damping, as was previously described. A Q of 0.5 is chosen, and the critical damping is determined by resistor $R_s + R_p$ in series with L_s .

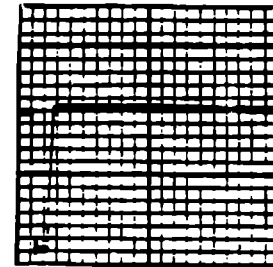
The other possibility for fall-time control is to use a low-Q inductive resistor connected to the DC charging supply. The same argument then applies here. That R that we looked at before can be an inductive resistor. A lumped inductor and a resistor can be used to make an inductive resistor. The relation for the damping time is given in Fig. 35 as:

$$\text{DAMPING TIME} = \frac{6}{\frac{R_s}{L_s} + \frac{1}{R_p C_s}}$$

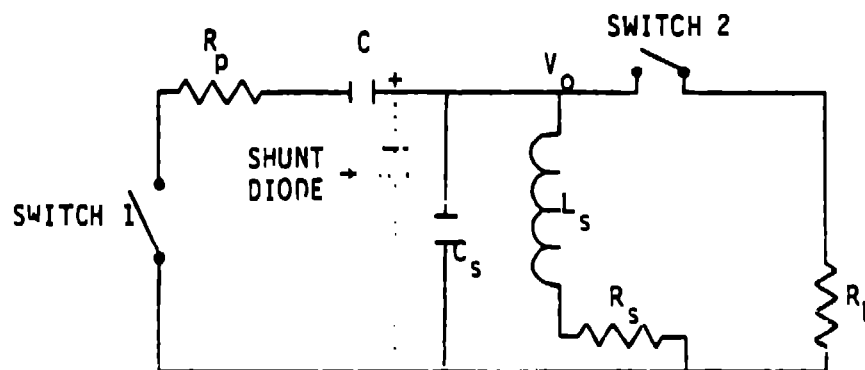
Experience has shown that one of the best wire-wound resistors available is made by Ward-Leonard if you can find a source of them. They take an unusually high surface stress. Rod Carborundum resistors are capable of absorbing even higher energies and are available for operation up to 100 kV in air (and considerably more in oil).



No shunt diode.



Good shunt diode.



1. CLOSE SWITCHES 1 AND 2.
2. AT THE END OF PULSETIME, SWITCH 1 IS OPENED.
3. VOLTAGE ACROSS R_L DROPS UNTIL SWITCH 2 OPENS
(E.G. LASER LOAD OR MAGNETRON)

THEN: SINCE CURRENT IS FLOWING IN L_s A SERIES RESONANT CIRCUIT OF C_s , R_s , L_s IS EXCITED:

OSCILLATIONS ARISE IF $\frac{R_s}{L_s} < \frac{2}{\sqrt{L_s C_s}}$ AT THE FREQUENCY $\omega^2 = \frac{1}{C_s L_s} - \frac{R_s^2}{4L_s^2}$

THESE OSCILLATIONS ARE DAMPENED AND APERIODIC IF

$$\frac{R_s}{L_s} > \frac{2}{\sqrt{L_s C_s}}$$

FOR EXAMPLE, IF $R_s = 50 \Omega$ $L_s = 500 \mu H$ $C_s = 50 \text{ PF}$

$$\frac{R_s}{L_s} = 10^5 \quad \frac{2}{L_s C_s} = 4 \times 10^8 \Rightarrow \text{OSCILLATIONS}$$

FIG. 35: INDUCTIVE RESISTOR USED TO CHARGE ENERGY STORAGE CAPACITOR AND PROVIDE FALLTIME DAMPING

AN INDUCTIVE RESISTOR WITH $\frac{L_S}{R_S} \approx \frac{1}{2} \sqrt{L_S C_S}$ CAN BE USED TO

TO OBTAIN AN APERIODIC FALLTIME $\propto e^{-At}$ THEN:

$$2A = \frac{R_S}{L_S} + \frac{1}{RC_S} \quad \text{AND} \quad \frac{1}{R} = \frac{1}{R_P} + \frac{1}{R_L}$$

GENERALLY, $R_P \ll R_L$, SO $R = R_P$ MEANING

$$2A \approx \frac{R_S}{L_S} + \frac{1}{R_P C_S}$$

NOTE: DAMPING IS THE SAME FORM FOR APERIODIC OR OSCILLATORY CASE. HERE THE VOLTAGE ≈ 0 IN A TIME $T_D = 3/A$ OR "DAMPING TIME"

T_D WHERE

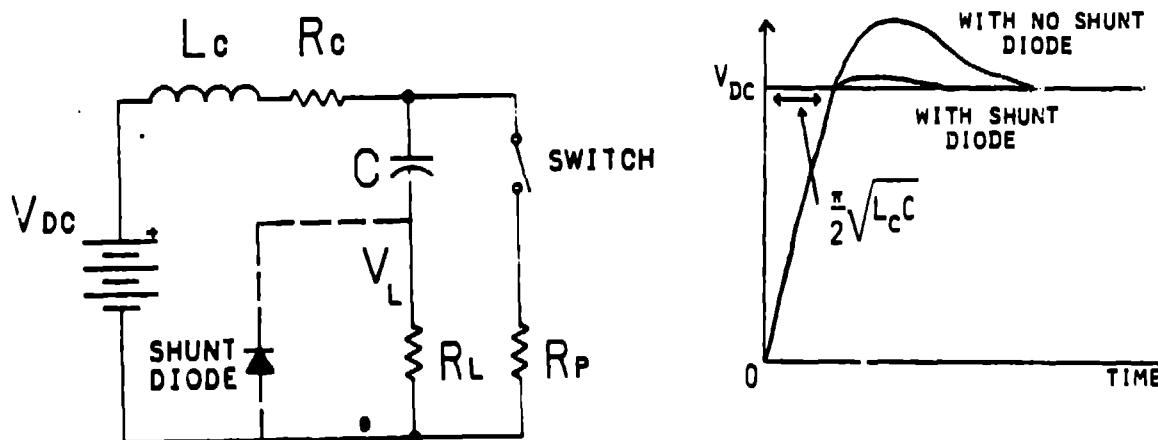
$$T_D = \frac{6}{\frac{R_S}{L_S} + \frac{1}{R_P C_S}}$$

Resistive charging is generally not used because the charging efficiency is only 50%, so inductive charging (Fig. 36) is required. If the damping components are chosen correctly, one can have critical damping. The capacitor C can be charged very quickly back to the DC voltage with a very small amount of overshoot. It can be shown that the overshoot on recharge is within 1% of the final voltage in the time $\pi/2 \sqrt{L_C C}$. If any significant overshoot on the pulse there is undesirable, a shunt diode can be put across R_L . It ensures that on the recharge, R_L doesn't carry any significant current. This is basically how high-efficiency, high-rep rate hard tube pulsers are charged. This L_C is chosen to recharge C in a fraction of the PRF period.

APPLICATION OF PULSE SHAPING NETWORKS TO HARD-TUBE PCS

This is included because it has a significant potential for constant-current drive circuits (Figs. 37 and 38). The same arguments apply to using a thyatron or any type of pulser switch imaginable. For the shunt LR network, if everything is correctly selected, and if $L = R^2 C$, it can be shown that the current under certain conditions remains relatively constant during the discharge interval. It's a great driver, for example, for flash tubes or for certain classes of lasers (not KrF lasers but a lot of other interesting ulow-mode lasers). There's one important thing here, which impacts on the design of modulators. The current in this network can remain monotonic (i.e., it won't reverse) for $(\pi/2) \sqrt{LC}$ seconds. This is an important result for solid-state diode recovery. The current should remain monotonic through a diode while it's recovering. If this condition is chosen during snubber design, the result is the correct element to put in series with the diode so the diode cannot suffer a very rapid current reversal. This is a vital element in a number of high-rep-rate modulator designs. Decoupling networks are needed to protect diodes in large modulators as without them oscillations can ensue. This network will hold the current constant for that period of time, $(\pi/2) \sqrt{LC}$. As long as this time is longer than the diode recovery

FIG. 36: INDUCTIVE CHARGING ELEMENTS



1. OPEN THE SWITCH.

RECHARGE IS CRITICALLY DAMPENED IF $Q = 0.5$, I.E.

$$\frac{\omega L_C}{R} = 0.5 \text{ AND } \omega^2 L_C \cdot C = 1 \text{ AND } R = \{R_C + R_L\}$$

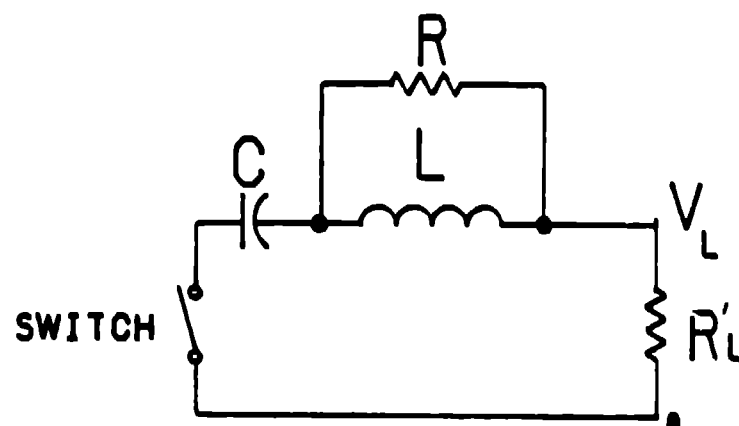
THEREFORE: $\frac{L_C}{\sqrt{L_C C} \cdot R} = \frac{1}{2} \quad \text{OR} \quad R = \frac{1}{2} \sqrt{\frac{L_C}{C}}$

IN THIS CASE, OVERSHOOT ON RECHARGE IS NEGLIGIBLE AND RECHARGE TIME TO WITHIN 1% OF FINAL VOLTAGE IS

$$\frac{\pi}{2} \sqrt{L_C C}$$

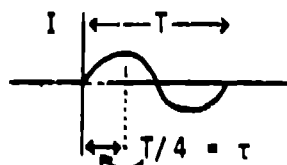
2. OSCILLATIONS MAY BE DAMPENED USING SHUNT DIODES ACROSS R_L TO CONTROL THE OVERSHOOT.

FIG. 37: APPLICATION OF PULSE SHAPING NETWORKS
TO HARD TUBE POWER CONDITIONING SYSTEMS



WHEN $L = R^2 C$, THE LOAD CURRENT HAS ZERO SLOPE AT $T = 0$,
IMPLYING A CONSTANT CURRENT SOURCE INDEPENDENT OF CIRCUIT Q.

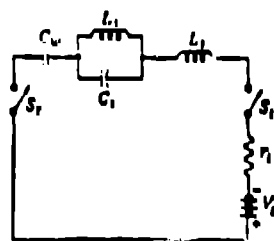
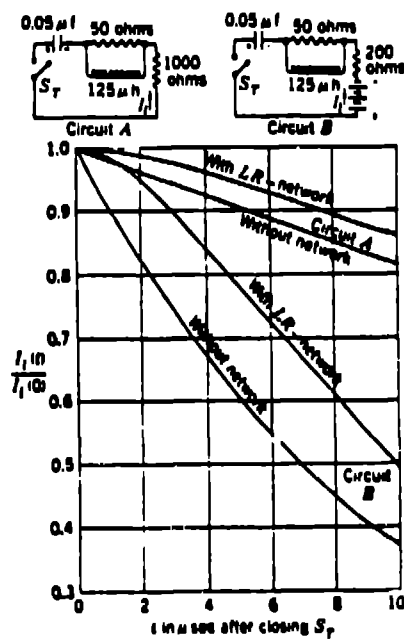
NOTE:



THE CURRENT REMAINS MONOTONIC FOR A TIME $\tau = \frac{\pi}{2} \sqrt{LC}$

IMPORTANT FOR SOLID STATE DIODE RECOVERY, SINCE CURRENT
FLOW THROUGH THEM MUST BE UNIDIRECTIONAL UNTIL THE DIODE
RECOVERS ($\approx 2-3 \mu s$).

FIG. 38: CONSTANT CURRENT PULSE SHAPING NETWORKS



Output circuit of a hard-tube pulser with a network of passive elements to maintain approximately constant load current during the pulse.

time, its precise value doesn't matter. If the diode can recover in three microseconds, the time is selected to be 5 to 10 microseconds. This is a very important point. One of these in series with diode bridge elements and recovery problems are negligible. One of these in series with the charging diode will keep the diodes from being damaged under any fault conditions except a prolonged DC fault. This network forces the current to remain monotonic during that short period of time.

CHARACTERISTICS OF SWITCH TUBES

A sketch of a vacuum switch tube is shown in Fig. 39. All of the rest of this discussion is based to a large extent on Tom Burkes' switch report. This sketch shows a profile of what tubes can be purchased (Fig 40), to about 400 kV can be bought at low peak currents. Three hundred megawatts peak power capability is available at just under 100 kV, although expensive to buy. Response time is generally in the area of microseconds because of the tube's size. The peak current is space-charge-limited, as it is cathode phenomena that limit all peak current.

Several points of interest are noted in Figs. 41-43.

1. Peak Current: Limited to 1-2 kiloamp by available cathode-emission structures. Improving cathode emission would make more peak current become available.
2. Lifetime observations: After 10,000 hours, the thoriated tungsten cathode emission drops. Oxide cathodes have shorter lifetimes than this by a factor of ten.
3. Anode voltage: The basic limit is vacuum breakdown from induced field emission.
4. X-Ray radiation: Especially troublesome for long pulse durations at high voltages and rep rates.

FIG. 39: SWITCH TUBE MATERIALS

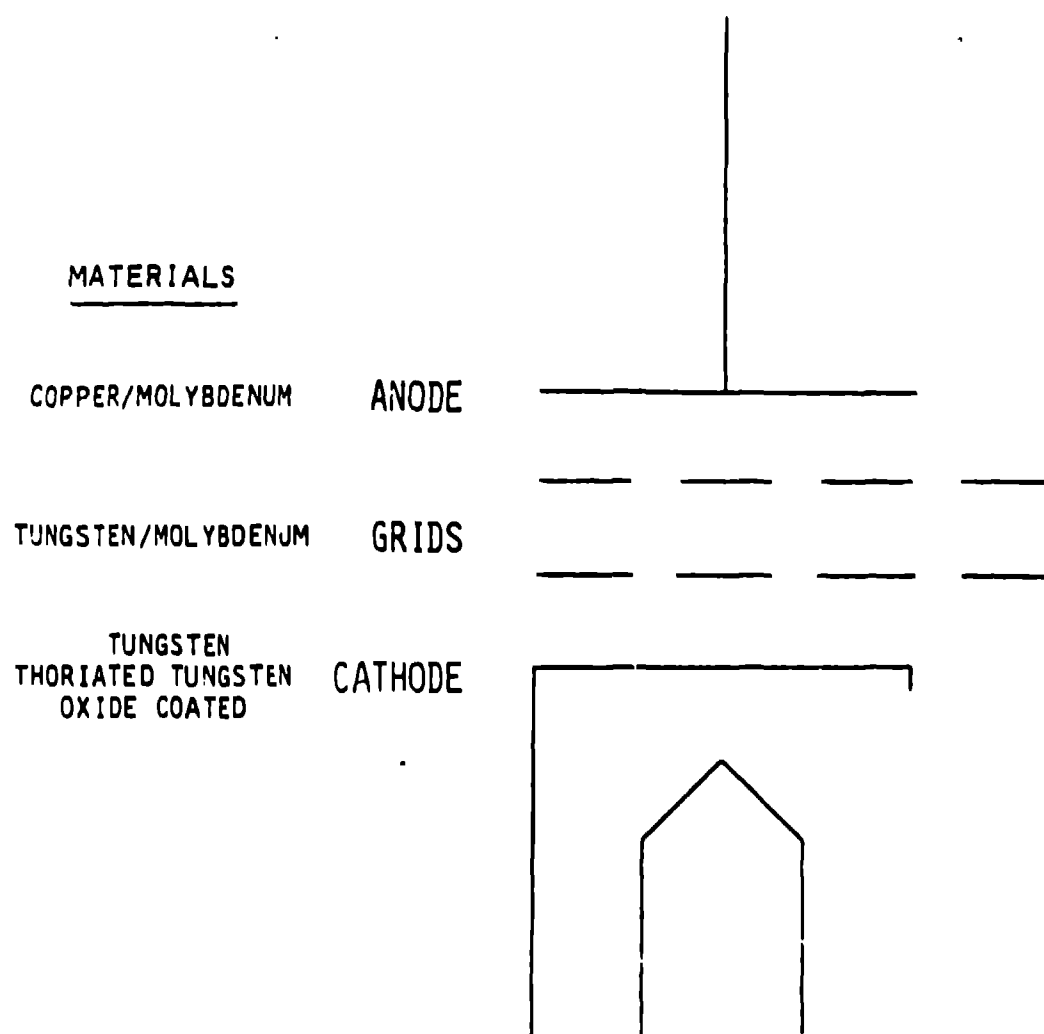
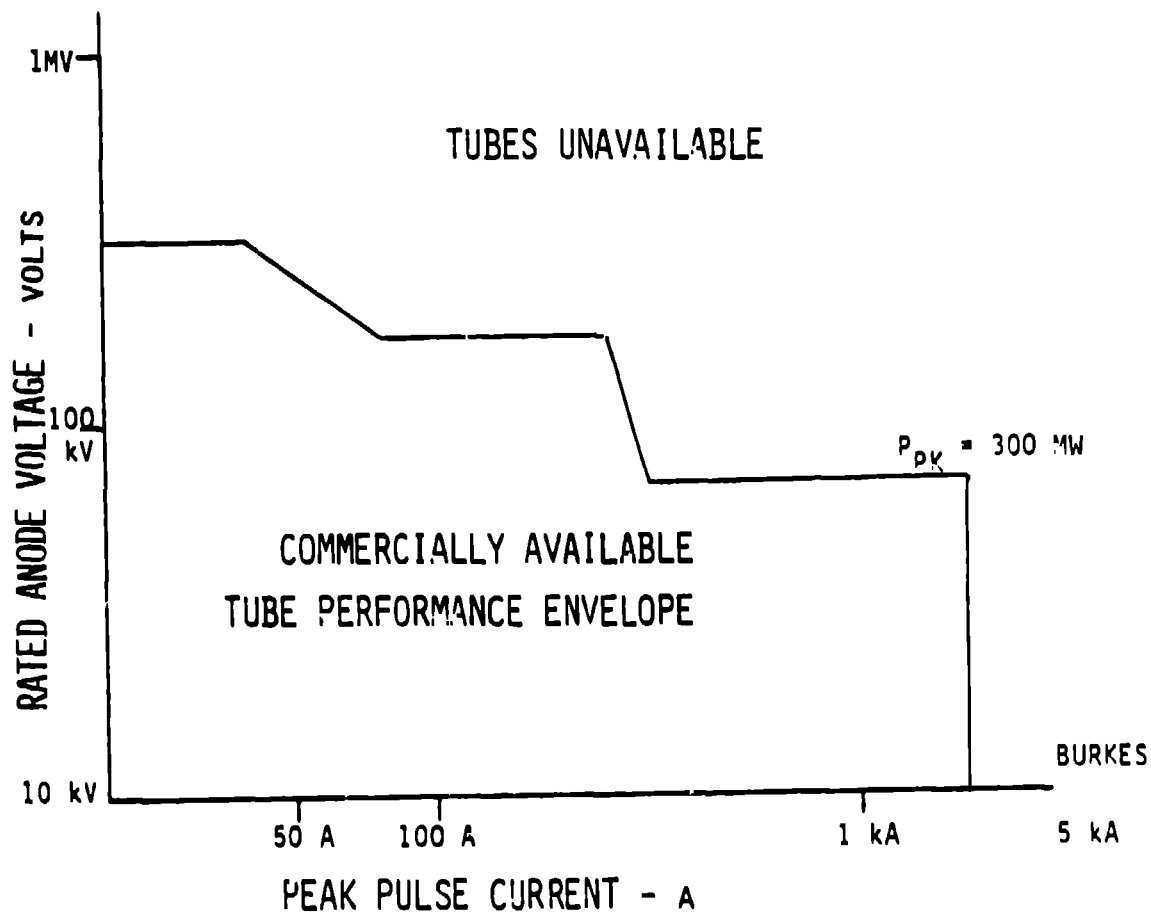


FIG. 40: CHARACTERISTICS OF HIGH VACUUM SWITCH TUBES



PEAK CURRENT IS SPACE CHARGE LIMITED.

FIG. 41: CHARACTERISTICS OF SWITCH TUBES

1. PEAK CURRENT: LIMITED BY CATHODE EMISSION AND GRID INTERCEPTION (AND CONSEQUENT HEATING) ≈ 1000 A.
2. LIFETIME: OXIDE CATHODES HAVE SHORT LIVES COMPARED TO THORIATED TUNGSTEN. I.E. 10,000 VS 1000 HOURS AS THE EMISSION DROPS.
3. ANODE VOLTAGE: LIMITED BY VACUUM BREAKDOWN FROM INDUCED FIELD EMISSION AND/OR POSITIVE ION BOMBARDMENT
 - SPACING/SURFACE EFFECTS
4. X-RAY RADIATION: A SIGNIFICANT PROBLEM AT HIGH REPETITION RATES, VOLTAGES, AND LONG PULSE DURATIONS.
 - ANODE PHENOMENON.
5. PRF: LIMITED PRIMARILY BY ANODE AND GRID HEATING:

$$I_{RMS}^2 = I_{PK}^2 (DF) \quad DF = \text{DUTY FACTOR}$$
6. CURRENT RATE OF RISE: IF C_s IS STRAY CAPACITANCE, THEN $R_s C_s$ IS FASTEST RISETIME WHERE R_s IS TUBE (SATURATED) ON RESISTANCE.

TUBE CHARACTERISTICS

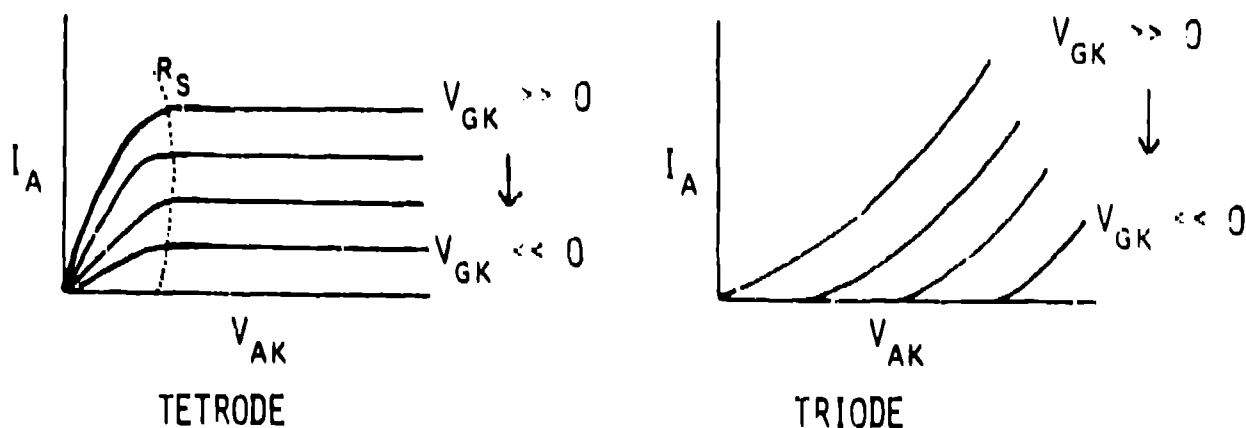
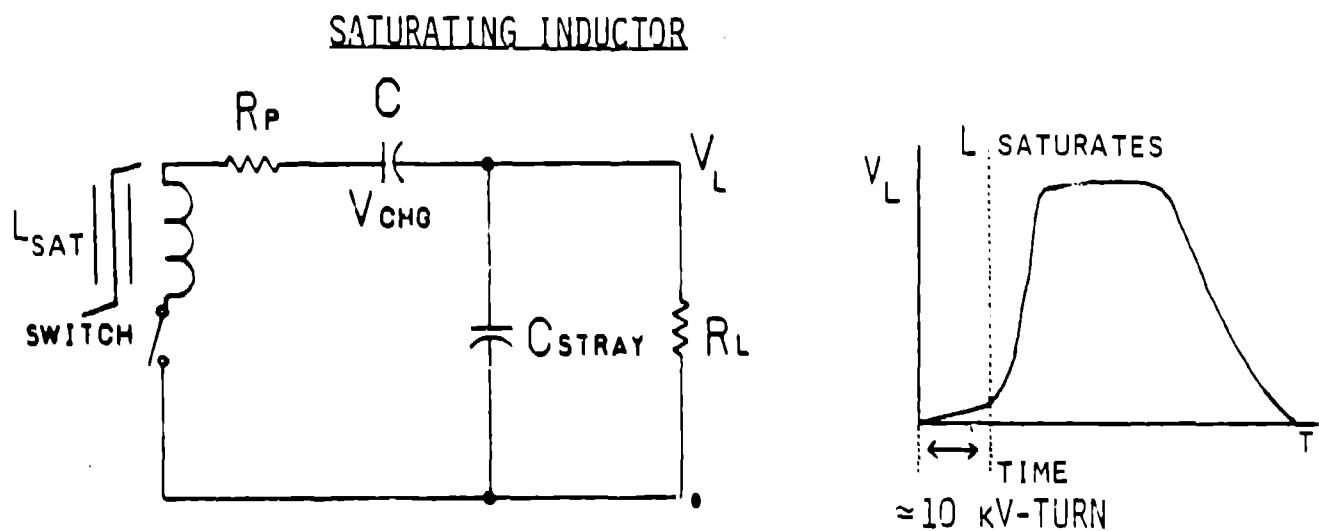


FIG. 42

6. RATE OF RISE OF CURRENT: NOTE THAT RESONANT CIRCUITS HAVE BEEN USED TO IMPROVE RISE TIMES, BUT GENERALLY AT THE EXPENSE OF AVAILABLE PEAK PULSE VOLTAGE



7. PULSE WIDTH: LIMITED BY GRID HEATING FOR ALL CATHODES; FOR OXIDE CATHODES SUBLIMATION OF OXIDE COATS GRIDS AND HENCE GRID EMISSION ENHANCED AT LOWER GRID OPERATING TEMPERATURES.
- CURRENT LIMITS ARE SEVERAL SECONDS ON AT $\approx 1 \text{ MW}$ LEVELS.

FIG. 43

8. DELAY: LIMITED BY LRC RESONANCE AND TRANSMISSION LINE EFFECTS IN THE CASE OF CABLE STORAGE . PFN SYSTEMS.
9. JITTER: USING DC ON HEATERS, JITTERS $\ll 1$ NS ACHIEVABLE.

5. PRF: Basically limited by anode and grid heating. The electrons are accelerated through the grids, each having a scattering cross-section, and some strike each grid, then the anode, heating them all up.
6. Current Rate of Rise: Fundamentally stray-capacitance and tube saturation-resistance limited. To increase this a saturating inductor can be used in the anode. Turning the tube on starts a little current flow through the large inductance. The ferrite then saturates, decreases the inductance, and a fast-rising output pulse is obtained. Unfortunately, when the tube is reopened, a long voltage decay ensues because the magnetic energy stored in the ferrite feeds energy back into the load.
7. Pulse Width Limit: Grid heating, a thermal effect limits the average power, except for oxide cathodes, where material sublimation limits temperatures because of enhanced field emission. Present limits are several seconds on time at the megawatt average power level.
8. Delay: Limited by RLC resonance effects in the external circuit and tube electron transit time.
9. Jitter: A couple of hundred picoseconds for planar triodes with DC heater power.

TYPICAL HARD-TUBE PCS

Last but not least, Fig. 44 shows what a hard-tube pulser looks like. Believe it or not they haven't fundamentally changed significantly since 1945. In this circuit there are a number of 6D21 tubes in parallel connected to a magnetron load through the two 0.01 microfarad storage capacitors. Turning the tubes on shorts the capacitors to ground, placing a negative pulse on the cathode of the magnetron. The pulse length is

FIG. 44: HIGH-POWER, SHORT-PULSE HARD-TUBE PULSER

The pulser was designed for a ...

maximum voltage-pulse amplitude of about 30 kv, and a pulse current of about 35 amp. Since it was desired that the short pulses be nearly rectangular, the output circuit had to be designed for a very high rate of rise of voltage. The inductance and the shunt capacitance in the pulser circuit therefore had to be kept as small as possible. In order to minimize the inductance introduced by the circuit connections, the components were mounted as close together as their physical size would permit, allowing for the spacing necessary to prevent high-voltage flash-over. The importance of keeping the stray capacitance small is evident when

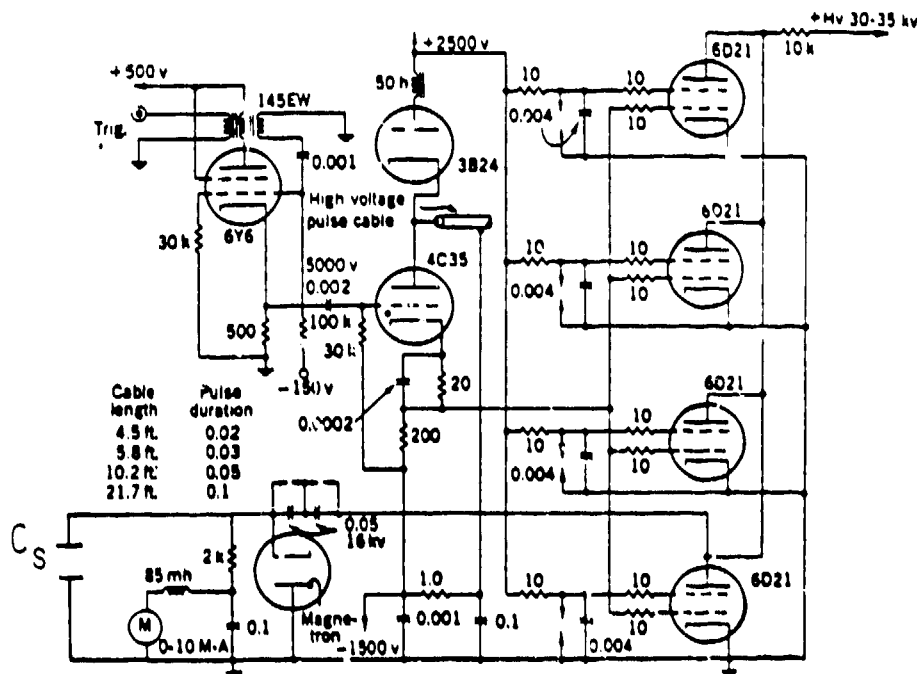


FIG. 8-12.—Schematic diagram of the circuit for a high-power short-pulse hard-tube pulser.

the charging current, $C, dv/dt$, for this capacitance is considered. For example, if the capacitance across the output terminals of the pulser is $20 \mu\text{f}$ and the voltage pulse at the load is to rise to 30 kv in $0.01 \mu\text{sec}$, the condenser-charging current is 60 amp. Since this current must flow through the switch tube, the maximum plate current needed to obtain a large value of dv/dt may be considerably greater than that needed to deliver the required pulse power to the load. The magnetron may have a capacitance of 10 to $15 \mu\text{f}$, between cathode and anode, so it is obvious that the additional capacitance introduced by the pulser circuit must be small.

6D21

CATHODE: Thoriated Tungsten, 150 Watts Heating Power
 $V_p(\text{max})$: 33 kv
 I_p
 $I_p(\text{max})$: 15 A